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BASES OF RADIO DIRECTION FINDING

by

I. S. Kukes, M. Ye. Starik

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U. S. BOARD ON GEOGRAPHIC NAMES TRANSLITERATION SYSTEM

Block	Italic	Transliteration	Block	Italic	Transliteration
А а	*А а*	A, a	Р р	*Р р*	R, r
Б б	*Б б*	B, b	С с	*С с*	S, s
В в	*В в*	V, v	Т т	*Т т*	T, t
Г г	*Г г*	G, g	У у	*У у*	U, u
Д д	*Д д*	D, d	Ф ф	*Ф ф*	F, f
Е е	*Е е*	Ye, ye; E, e*	Х х	*Х х*	Kh, kh
Ж ж	*Ж ж*	Zh, zh	Ц ц	*Ц ц*	Ts, ts
З з	*З з*	Z, z	Ч ч	*Ч ч*	Ch, ch
И и	*И и*	I, i	Ш ш	*Ш ш*	Sh, sh
Й й	*Й й*	Y, y	Щ щ	*Щ щ*	Shch, shch
К к	*К к*	K, k	Ъ ъ	*Ъ ъ*	"
Л л	*Л л*	L, l	Ы ы	*Ы ы*	Y, y
М м	*М м*	M, m	Ь ь	*Ь ь*	'
Н н	*Н н*	N, n	Э э	*Э э*	E, e
О о	*О о*	O, o	Ю ю	*Ю ю*	Yu, yu
П п	*П п*	P, p	Я я	*Я я*	Ya, ya

*<u>ye</u> initially, after vowels, and after ъ, ь; <u>e</u> elsewhere.
When written as ё in Russian, transliterate as yё or ё.

RUSSIAN AND ENGLISH TRIGONOMETRIC FUNCTIONS

Russian	English	Russian	English	Russian	English
sin	sin	sh	sinh	arc sh	$\sinh^{-1}$
cos	cos	ch	cosh	arc ch	$\cosh^{-1}$
tg	tan	th	tanh	arc th	$\tanh^{-1}$
ctg	cot	cth	coth	arc cth	$\coth^{-1}$
sec	sec	sch	sech	arc sch	sech^{-1}
cosec	csc	csch	csch	arc csch	csch^{-1}

Russian	English
rot	curl
lg	log

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Chapter 6.

Errors of radio direction finder, connected with radiowave propagation.

6.1. Effect of the abnormal polarization of electric field.

To external equipment/device of radio direction finder besides the terrestrial normal-polarized wave, can approach also the waves, reflected from upper air. In the general case the electric field of the wave reflected contains the vertical and horizontally polarized components, moreover dimensional orientation of the components of electric field, and also the relationship/ratio of their amplitudes and phases change in time.

The radiation patterns used at present in the radio direction finders of antenna systems in radiowave propagation at an angle to the horizon usually do not coincide for vertical and horizontal components electric field. Therefore the resulting radiation pattern

for a total field does not coincide with radiation patterns for field component. We see earlier that with amplitude direction-finding method, utilizing an antenna radiation pattern of the system of radio direction finder for a vertical (or horizontal) electric field, it is possible to determine correct bearing on radio station. From that which was presented it follows that during the determination of bearing from the resulting radiation pattern they are obtained in the general case of error. Since the parameters of the reflecting ionized layers of the atmosphere continuously change, changes the polarization of radio waves, and also error in time. They are called polarizational errors. In phase direction finders the simultaneous reception/procedure of both components of electric field also is led to errors.

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As is known, on ultra short waves under normal conditions, does not occur the reflection from the ionized layers. On middle and long waves the reflection from upper air is observed mainly by night; therefore on these waves polarization errors are developed to the most powerful degree by the night and they are frequently called in radio direction finders "night effects". On short waves the reflection from upper air occurs during all days.

Since the presence of polarizational errors is not connected with a change in the direction of propagation, the polarizational errors of radio direction finder one should relate to tool houses.

As the basis of the elimination of polarizational errors, can be placed one of the following principles:

1) The use of the antenna of the system at whose radiation patterns for vertical and horizontal electric fields would coincide completely or at least in that part from which is determined the bearing (for example, application/use by the antenna of system with the diverse framework);

2) decrease down to the minimum limit of reception/procedure of one of the components electric field, usually horizontal (system with the spaced antennas, the compensated for framework and so forth);

3) the elimination of the reception/procedure of the wave reflected and the preservation/retention/maintaining of reception/procedure one of terrestrial normal-polarized field (the sampled-data systems of direction finding).

Polarizational errors are the rapidly changing random errors and their average value with sufficient time of direction finding is

close to zero.

In order to decrease the manifestation of these errors, usually in the radio direction finders, constructed according to the second principle, are taken with direction finding several readings and is designed average bearing.

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6.2. Determination of error due to the abnormal polarization of electromagnetic field.

Let us examine action on the radio direction finder of terrestrial and reflected radio waves.

Let the electric field of terrestrial wave has elliptical polarization and it it is possible to decompose on E_1 - vertical component and $E_2e^{i\varphi_1}$ [1] - horizontal component in the plane of propagation.

FOOTNOTE [1]. Here and throughout phase is counted off relative to E_1.

ENDFOOTNOTE.

The horizontal component of terrestrial wave in the plane, perpendicular to direction of propagation, at a great distance from transmitter attenuates and it is equal to zero. The decrease in the horizontal field the greater, the greater the equivalent ground conductivity.

The electric field from upper ionospheric layers of wave reflected contains components: $E_{01}e^{i\varphi_0}$ - vertical field; $E_{02}e^{i\varphi_{01}}$ - horizontal field in the plane of propagation; $E_{03}e^{i\varphi_{03}}$ is a horizontal field in the plane, perpendicular to the plane of propagation.

In the case of the linear polarization of the electric field of the wave reflected, which has amplitude E_0, the phase Φ_0, the angle of polarization γ and the angle of the slope of a front of wave β, its components will be: $E_0 \cos\gamma \cos\beta e^{i\varphi_0}$ - a vertical field; $E_0 \cos\gamma \sin\beta e^{i\varphi_0}$ - a horizontal field in the plane of propagation; $E_0 \sin\gamma e^{i\varphi_0}$ - a horizontal field in the plane, perpendicular to the plane of propagation.

Let us designate: $H_1F_1(\alpha, 0, \beta)$ - effective height and

directional characteristic by the antenna of system for components the electric field of wave in the plane of propagation: $d_2 F_2(\alpha, \theta, \beta)$ - the same characteristics for a component electric field in the plane, perpendicular to direction of propagation.

Sometimes for in the antenna of the system of radio direction finder directional characteristic $F(\alpha, \theta, \beta)$ can be presented in the form of the product of two characteristics: $\Phi(\alpha, \theta)$ - in horizontal plane and $f(\beta)$ - in vertical plane, i.e.,

$$F(\alpha, \beta, \theta) = \Phi(\alpha, \theta) f(\beta),$$

where θ is an angle of the direction of the oriented radio station with the initial reference line of bearing; α is an angle of the orientation of directional characteristic of relatively initial reference line.

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Let us write condition for the reading of bearing in the general case:

$$\begin{aligned} G\{&E_1 H_1 F_1(\alpha, \theta, \beta = 0) + E_2 e^{j\varphi_2} H_2 F_1(\alpha, \theta, \beta = 90^\circ) + \\ &+ E_{01} e^{j\varphi_{01}}[1 + R_1 e^{j(\psi_1 + \eta_1)}] H_1 F_1(\alpha, \theta, \beta) + \\ &+ E_{02} e^{j\varphi_{02}}[1 + R_1 e^{j(\psi_1 + \eta_2)}] H_2 F_1(\alpha, \theta, \beta = 90^\circ) + \\ &+ E_{03} e^{j\varphi_{03}}[1 + R_2 e^{j(\psi_2 + \eta_2)}] H_2 F_2(\alpha, \theta, \beta)\} = g^*, \end{aligned} \quad (6.1)$$

1.

FOOTNOTE [1]. In the case of propagation of VHF on top, for example from aircraft, $E_1 = E_2 = 0$ and E_{01}, E_{02}, E_{03} in different relationship/ratios are emitted by the antenna of aircraft. ENDFOOTNOTE.

where R_1, ψ_1 are the module and the argument of the coefficient of terrain echo of the vertically polarized field; R_2, ψ_2 [- the same for horizontal-polarized field η_1, η_2 is the delaying phase earth-reflected wave relatively falling directly for vertical and horizontal fields; G is an operation of determining the bearing (obtaining the maximum, minimum and so forth); g is the required for obtaining bearing result (zero, the maximum value of lug/lobe, etc.).

From Fig. 6.1 it follows that if sensors are arrange/located at height/altitude h_n, then η in (6.1) will be (see Fig. 6.1)

$$\eta = \frac{2\pi}{\lambda}(AB + BC) = \frac{2\pi}{\lambda} AD = \frac{4\pi}{\lambda} h_n \sin\beta,$$

where AB and BC is the increment of the ray/beam, reflected from the earth/ground, D - mirror image C and BC = BD.

For a goniometric system from n of the vertical wire antennas when the separation of antennas is much less than the wavelength, we have:

$$F_1(\alpha, \theta, \beta) = \sin(\alpha - \theta)\cos\beta,$$
$$F_2(\alpha, \theta, \beta) = \cos(\alpha - \theta),$$
$$H_1 = \frac{2\pi b h_{\text{..}} N}{\lambda}.$$

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H_2 depends on the diagram of connection of antennas (U-, H-shaped, etc.). If we substitute these values in (6.1), then by formulas (III.3)-(III.7) can be calculated $a_{\text{мин}}$ and bearing error $\Delta\theta_n$, and also ratio A/B. Error increases with an increase in ratios H_2/H_1 and E_{03}/E_{01} (with an increase γ), and also with an increase of angle β.

For the characteristic of the degree of susceptibility by the antenna of the system of amplitude radio direction finder to errors due to abnormal polarization is introduced the concept of the error of standard wave - $\Delta\theta_{\text{ст}}$. This is that polarizational error which appears with the direction finding of one abnormal and plane-polarized wave reflected when $\gamma = \beta = 45°$.

For a frame radio direction finder with the reading of bearing on zero audibility.

$$H_1 = H_2,\ F_1(\alpha, \theta, \beta) = \sin(\alpha - \theta),$$
$$F_2(\alpha, \theta, \beta) = \cos(\alpha - \theta)\sin\beta.$$

Set/assuming in (6.1)

$$E_1 = E_2 = E_{02} = 0,\ E_{01} = E_0 \cos\gamma,\ E_{02} = E_0 \sin\gamma,\ \varphi_{01} = \varphi_{02} = 0,$$

from condition g = 0 we will obtain for the polarizational error

$$\operatorname{tg} \Delta\theta_n = \operatorname{tg}(\alpha - \theta) = \left| \operatorname{tg}\gamma \sin\beta \frac{1 + R_2 e^{j(\psi_2 + \gamma_{2})}}{1 + R_1 e^{j(\psi_1 + \gamma_1)}} \right|.$$

If we at an angle γ understand the angle of the polarization of the resulting wave taking into account terrain echo and to suppose that the phases of the vertical and horizontal components of electric field coincide, then we will obtain

$$\operatorname{tg} \Delta\theta_n = \operatorname{tg}\gamma \sin\beta. \qquad (6.2)$$

This expression is obtained into §3.3. The error of the standard wave of the framework at $\gamma = \beta = 45°$ is determined from the condition

$$\operatorname{tg} \Delta\theta_{cr} = 0{,}707 \text{ or } \Delta\theta_{cr} = 35{,}3°.$$

The methods of the practical determination of polarizational errors are given in chapter 9.

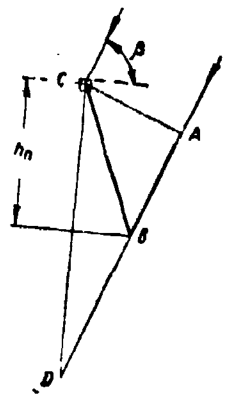

Fig. 6.1. Course of ray of the wave reflected.

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6.3. Systems of radio direction finders, free from polarizational errors.

System with the diverse framework.

The described (into §3.7) system with two diverse framework is almost completely free from polarizational errors, since the

horizontal component of the field reflected is not created errors with direction finding.

Main disadvantages in the systems with the diverse framework they are:

1. Sharp decrease in the sensitivity of system with the elongation of wave.

2. Complexity of engineering goniometric system with diverse framework. goniometric system of two pairs of the diverse framework whose planes must be parallel, has different sensitivity for the different angles of arrival of wave. Sensitivity is equal to zero for direction of propagation, perpendicular to the planes of the framework.

Is necessary instead of each framework to take two mutually perpendicular changed over framework (in all are obtained eight frameworks); as a result the system strongly becomes complicated. At the same time, rotary system possesses large inertia, which increases time of the taking of bearing.

3. Presence in radiation pattern besides two zeros, determining accurate bearing, an additional two false minimums of audibility of

signal. It should be noted that the last/latter deficiency/lack is not virtually essential, since "the false" minimums with the direction finding of the waves reflected are strongly are blunted and slightly differed from the true.

Furthermore, is required large thoroughness in construction and in the production of system with the diverse framework; thus, for instance, it is necessary to fulfill with high accuracy the parallelism of the planes of the framework (0.1-0.2°), the identity of the distances of the framework of receiver, etc.

Systems with the spaced antennas. Different connection diagrams.

The operating principle of such antennas is described earlier (§§2.2, 3.6, 3.10. Antennas, are constructed so that the external system would be accepted only vertical component of electric field.

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The exception/elimination of the reception/procedure of horizontal component is led to the elimination of polarizational errors. Sometimes on ultra short waves is realized the reception/procedure of

one horizontal component of field.

The simplest system is the so-called H-shaped system (H-shaped system). The pattern of the connection of the vertical wire antennas in H-shaped system is shown in Fig. 6.2.

Emf from each of the vertical wire antennas will be feed/conducted to coils K by the horizontal pairs of the wires of identical size/dimension (bc = cd = hg = ge), arrange/located at very close distance from each other. As a result of the complete identity of wires in them, they are induced under the effect of horizontal electrical field of emf of identical value and identical phase. Difference these emf, forming voltage on coil K, is equal to zero, so that on system operates only vertical component of electric field. Therefore polarizational errors must be absent.

In order to ensure the absence of reception/procedure to the horizontal parts of the antenna, are necessary not only the equality of the lengths of horizontal wires, but also identical loads at their end/leads. The load of horizontal wires are the upper and lower halfdipoles ab, ae and HF, df whose parameters in the general case are different due to their location at the different height/altitude above the earth/ground.

On short and ultra short waves sometimes for the equation of loads from halfdipoles they raise above the earth/ground entire antenna system. It is possible lower halfdipoles to take somewhat of the smaller length than upper, but by these they steady from upper and lower halfdipoles only in narrow frequency band.

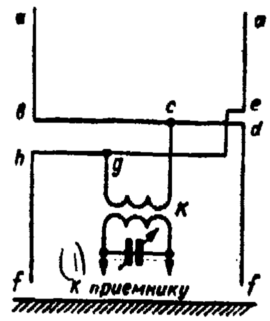

Fig. 6.2. H-shaped system of spaced antennas.

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Therefore this method of a decrease in the polarizational errors must not consiler acceptable for a wide-range radio direction finder. Furthermore, the optimum difference in the lengths of upper and lower halfdipoles depends on the parameters of soil, i.e., it is variable for the different building grounds of radio direction finler. The nearer the parameters of soil to dielectric, the lesser its effect on asymmetry.

Other reasons for the polarizational errors in H-shaped system they are:

a) terrain echo of the field, created by current in the screen of feede and the induction to them emf in vertical condutors.

b) the presence of the communication/connection between the wires of feeder and input circuit of receiver.

The analysis of the components of the polarizational error of H-shaped system is given in [6.5].

H-shaped system can be applied in the form of the rotary pair of antennas. In its this form frequently they utilize on short and ultra short waves. Antenna system rotates around motionless receiver or together with it.

It is possible to also apply goniometric H-shaped system. Goniometer with receiver in this case they place in cabin, which is located in the center of antennas.

On long waves the lower parts of the vertical wire antennas (df and ~~HF~~ hf in Fig. 6.2) are taken usually shorter than upper, and for the balancing of the load of horizontal wires in lower wires, are connected the equalizing capacitance/capacities C (Fig. 6.3). In this form H-shaped system is called that which was balanced (or equilibrium). To fit capacitance/capacity C so that they were

equivalent to the load from upper dipoles on large wave band, is difficult. Balancing of H-shaped system is applied therefore with the limited wave band.

Are known the cases of applying the balanced H-shaped system, also, on short waves.

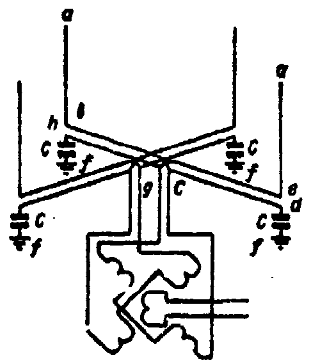

Fig. 6.3. Balanced H-shaped system.

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In VHF range sometimes is realized antenna system with the dipoles whose orientation can be changed for the reception/procedure either of horizontal or vertical electric field (Fig. 3.51).

The error of the standard wave of the virtually realized H-shaped systems lie/rests within limits [1.10]:

at frequencies, close to 10 MHz, 4°,

at frequencies, close to 1 MHz, 3°,

at frequencies, close to 0.3 MHz, 4-8°.

Widest use will receive besides the described two systems also a U-shaped system and system with conversion transformers.

Operating principle U-shaped of system is shown in Fig. 6.4. The reception/procedure of the horizontally polarized component of the electric field of electromagnetic waves they strive as far as possible to decrease. For this, are placed the horizontal parts of the antennas (feeders) at certain depth on the order of 1.5-2 m, where the horizontal field is attenuate/weakened [1], and also is applied the system of the grounding of the shell of feeders.

FOOTNOTE [1]. Feeders are placed below depth of soil freezing where the parameters of soil are more constant during one year. ENDFOOTNOTE.

The degree of the weakening of the horizontal intensity of field at depth 2 m due to absorption in the earth/ground is shown in Fig. 6.5 for the case of two soils:

$\varepsilon = 10,\ \sigma = 10^{-2}\ \frac{1}{(1)\text{ом}\cdot\text{м}}$ ((2)рис. 6.5,б) — (3)влажная почва,

$\varepsilon = 5,\ \sigma = 10^{-3}\ \frac{1}{(1)\text{ом}\cdot\text{м}}$ ((2)рис. 6.5,а) — (4)сухая почва.

Key: (1). Ω•m. (2). Fig. (3). humid soil. (4). dry soil.

From curve/graphs it follows that at depth 2 m with $\sigma = 10^{-3}$ 1/Ω•m the minimum weakening of the horizontal component in the range of waves 20-3000 m ranges from 2 to 7.5 times. With $\sigma = 10^{-2}$ 1/Ω•m it changes from 8 to 16 times.

Thus, with ground conductivities less than 10^{-2} 1/Ω•m, especially on short waves, weakening is very small and one fill-up of feeders into the earth/ground operates not very effectively.

Fig. 6.4. П-shaped system.

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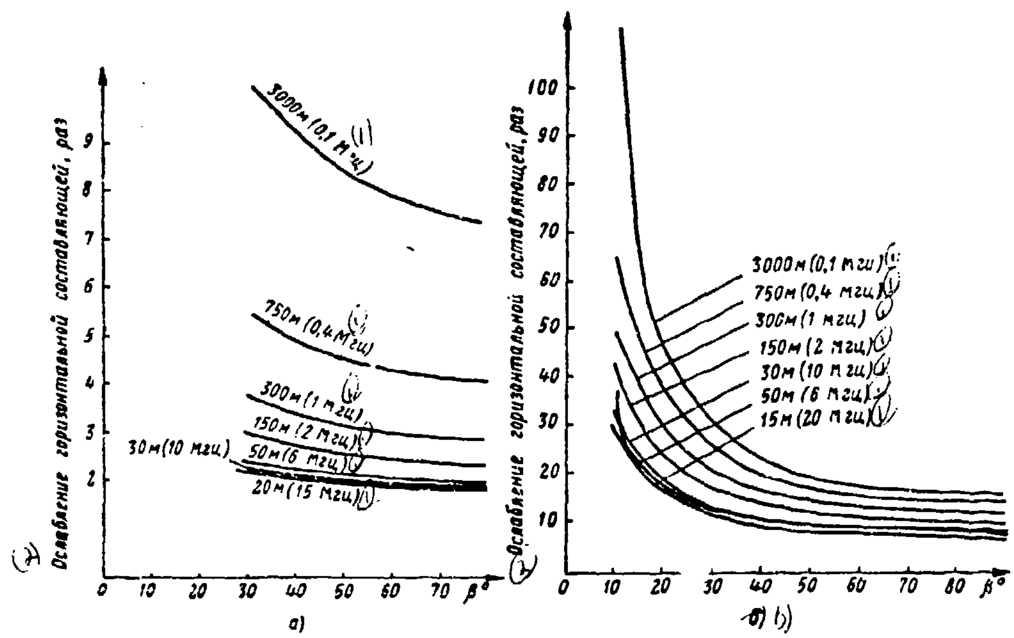

Fig. 6.5. Weakening of the horizontal component of field at depth 2 m: $a - \varepsilon = 5,\ \sigma = 10^{-3}\ \frac{1}{ом \cdot м}$; $b - \varepsilon = 10,\ \sigma = 10^{-2}\ \frac{1}{ом \cdot м}$.

Key: (1). MHz. (2). Weakening by horizontal component, times.

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If horizontal electric field excites in the shell of feeder certain voltage/stress E_{o6}, then by applying careful grounding of the end/leads of the shell of feeder it is necessary to attain a maximum decrease in this voltage/stress on the end/leads of the shell and in feeder.

The harmful voltage/stress, transmitted from shell into feeder, will be

$$E_{\Phi}=\left|kE_{o6}\frac{Z_{3}}{Z_{3}+Z_{o6}}\right|,$$

where Z_{3} is resistor/resistance of grounding of the end/lead of the shell of feeder; Z_{o6} - the resistor/resistance of the half of the length of shell, $Z_{o6}=j\rho_{0}\operatorname{tg}m_{0}l_{0}$; m_{0}, ρ_{0} - propagation constant and the wave impedance of shell; l_{0} is a half of the length of shell; k - the coefficient of the transmission of the voltage/stress from the end/lead of the shell into the feeder of antenna.

Than is less $\frac{Z_{3}}{Z_{o6}}$, i.e., how carefully is done grounding, those

less E_ϕ and is better the polarizational protection of U-shaped system.

Besides a good grounding under antennas for a decrease in the harmful effect of horizontal field, is applied the supplementary long wire, connected with the shell of feeder and packed in the earth/ground its as continuation (Fig. 6.6). By this is reached the symmetrical location of the vertical wire antenna of relatively secondary horizontal field and decreases the action of this field on antenna. The extension of the wire can be grounded.

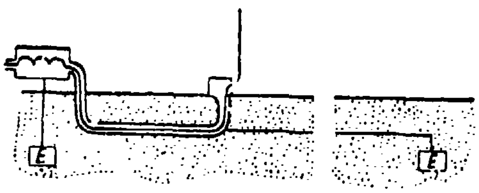

Fig. 6.6. Supplementary elongation of feeder.

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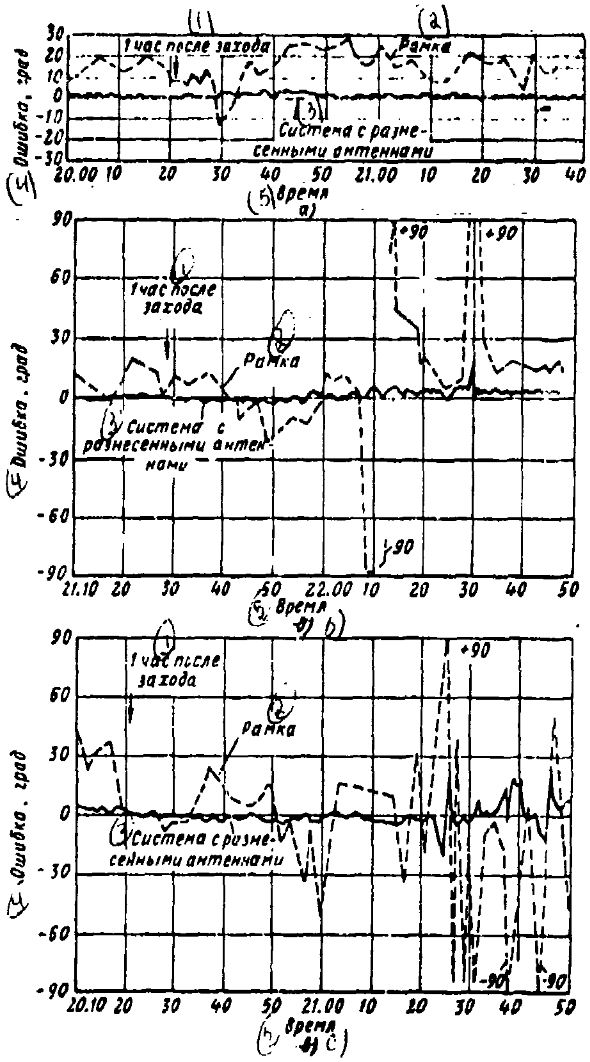

Fig. 6.7. Comparison of results of direction finding by framework and U-shaped system: a) good results; b) average results; c) poor results.

Key: (1). hour after approach. (2). Framework. (3). System with the spaced antennas. (4). Error, deg. (5). Time.

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Depending on the conditions of propagation, the quality of the work of U-shaped system can somewhat change. Figures 6.7 gives good, average and poor results of the work of U-shaped system [1.10].

U-shaped system with the buried into the earth/ground feeders is applied in places with large ground conductivity, when $\sigma \geqslant 10^{-2}$ 1/Ω•m. With such soils is observed the noticeable weakening of horizontal field at depth 1.5-2 m and, furthermore, sufficiently is easily attained a good grounding of shells of feeders.

For the best protection from the reception/procedure of horizontal field, it is possible the feeders of U-shaped system to place above or under counterweight in the form of wire gauze. A radius of counterweight for each antenna is designed from that consideration so that not less than 90o/o power of radiation current would be closed to counterweight. This frequently is led to the fact

that a radius of counterweight is obtained the equal to three-four heights of antennas. Work [6.7] gives the calculation of the optimum number of wires of grid during their laying along radii under each antenna. The pattern of the fulfillment of the grid of radio direction finder is shown in Fig. 6.8. Grid consists of square cells with side d.

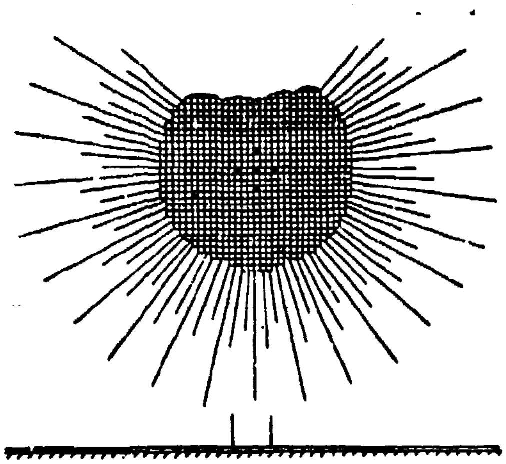

Fig. 6.8. Schematic form of grid.

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The separation can be calculated from the condition so that the overall length of wires during parallel laying was equal to the length of wires with their radial location. If according to calculation according to [6.7] radius of grounding under each antenna it is equal to r, the number of wires n, then the separation in mesh is obtained equal by d $\approx$ πr/n.

Grid usually assembles circular shape, with radius R = b + r (2b

- the separation of antennas). On the perimeter of grid, are soldered the lengthening wires whose length one should fit so that would be realized the effective grounding on the perimeter of grid on all frequency band of the radio direction finder, i.e., resistor/resistance earth referenced of the soldered to the perimeter of grid wire must be equal to zero. For this, along the length of wires, must be placed the odd number of quarter wavelengths of radio direction finder (taking into account the shortening of wave in soil). The total number of wires is selected order 100 with the alternating length, calculated for the different waves of range.

Figures 6.9 shows results of use of wire gauze with soil by conductivity $\sigma = 10^{-3}$ 1/Ω•m. Diameter of grid 31 m, the size/dimensions of mesh 0.6 x 0.6 m. Is investigated radio direction finder with 4 spaced antennas; the height of antennas 7.3 m, separation 7.3 m. Feeders are placed to the earth directly under grid.

In Fig. 6.9 are plotted the polarizational errors which are determined at frequencies 3-9 MHz with the aid of the heterodyne, raised, so that angle $\beta = 11°$.

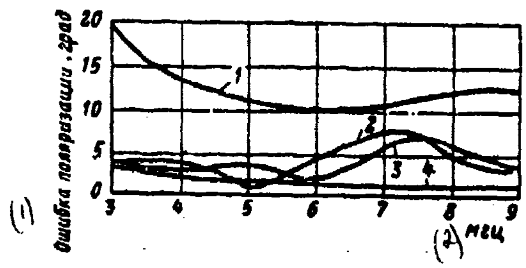

Fig. 6.9. Polarizational errors of U-shaped system with grid.

Key: (1). Error of polarization, deg. (2). MHz.

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Angle γ it composes 75°. In curve 1 are plotted/applied the polarizational errors of system without grid, in curve 2 - with grid with the soldered to its edges 36 groups of wires, each group consisting of three wires in consecutively changing 20, 15, 11 m length (a total of of 108 wires). Curve 3 gives errors with the same grid with 36 groups of wires, each of two wires with a length of 15 m even 25 m (72 wires). In curve 4 for a comparison are shown the polarizational errors of the same direction finder during installation on soil with good conductivity $\sigma = 3 \cdot 10^{-2}$ 1/Ω•m and with buried on depth 1.5 m feeders [6.4].

On the basis of available material, it is possible to arrive at the conclusion that during the application/use of a metallization of the surface of soil it is possible to establish/install U-shaped system in soils with the average conductivity, greater than 10^{-3} $1/\Omega{\bullet}m$. In ground conductivities 10^{-3} $1/\Omega{\bullet}m$ and less one should apply H-shaped system.

In [6.8], is described the simplified method of the metallization of the earth/ground in by four-antenna to U-shaped system in the range of frequencies 1-7 MHz, which consists in laying above each feeder of the antenna not of circular grid, but metallic band meshes.

Let us note the difference in the method of the elimination of the reception/procedure of the horizontal component of electric field in U-shaped and H-shaped systems. In U-shaped system this elimination is based on the screening of horizontal feeder. In H-shaped system the reception/procedure of horizontal field is eliminated as a result of mutual compensation; the loads of both wires of horizontal feeder must be completely identical, i.e., system must be symmetrical, which is virtually connected with those determined to difficulties.

In system with conversion transformers (Fig. 6.10) is placed the problem to facilitate the achievement of the symmetric loading of two wires of horizontal feeders. The lengths of the halves of the vertical wire antennas are different. In order to eliminate the effect of the dissimilarity of loads from these antennas on the horizontal wires, the latter are included through conversion transformers with small capacitance/capacities between windings.

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The lesser these capacitance/capacities, the lesser is developed the asymmetry of vertical condutors and the lesser the error of radio direction finder due to the reception/procedure of horizontal electric field.

It is possible in the lower halves vertical dipoles to include/connect compensating chains (on the long waves of capacitance/capacity). Then is obtained system transformer balanced.

Is applied also transformer system with the grounded vertical wire antennas. In this system for a decrease in the reception/procedure of the horizontal component of electric field, is required besides small capacitance/capacities between the windings of transformer the even better grounding of the vertical wire antennas.

As will be shown that the characteristic of susceptibility by the antenna of system polarizational to errors was introduced the error of standard wave.

In the U-shaped system of radio direction finder and in other systems whose feelers are placed in the earth/ground or it is close above the earth/ground at height/altitude less $\frac{\lambda_{мин}}{4}$ and in presence of one coming in from above electromagnetic field, the polarizational error is completely determined by the angle of the slope of a front of wave β and polarization γ of electrical field component. These systems they characterize by the error of standard wave.

For the raised H-shaped system, especially on the ultra short waves where delivery head can reach several wavelengths, the differences of the phases between that which falls and that which was reflected from the earth's surface (or sea) the components of electric field for vertical and horizontal polarizations they can be different. Because of this to the maximum of total field (directly coming in and reflected from the earth/ground) for horizontal polarization can correspond the minimum of total field for vertical polarization and vice versa.

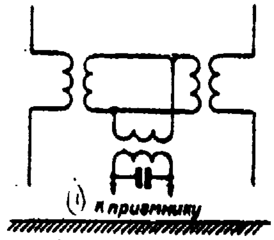

Fig. 6.10. Antenna circuit with transformers.

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Thus, at the point of direction finding considerably changes the relationship/ratio of horizontal and vertical electric fields. The polarization of the resulting field depends on distance of transmitter and delivery head by the antenna of the system of radio direction finder. This can lead to appearance at short distances from the transmitter of large polarizational errors (the horizontal component of electric field it is great, in vertical close to zero). At large distances from transmitter, when the angle of incidence of wave front becomes more than the angle of slip of Brewster, phase of reflection coefficients for a vertical and horizontal electric field they become identical and the relationship/ratio between vertical and horizontal radiation fields is retained during change in altitude.

So [6.2], at frequency 130 MHz, during the arrangement/permutation of H-shaped system at height/altitude 4λ and during propagation above humid soil ($\sigma = 10^{-2}$ $1/\Omega \bullet m$, $\varepsilon = 10$) Brewster angle is equal to 20° and with the direction finding of aircraft which flies at height/altitude 1500 m, the described phenomenon occurs at distances to 4.5 km; above sea the angle of slip of Brewster is equal to 84° and this distance reaches to 16 km. At large distances the real relationship/ratio between the vertical and horizontal components of electric field is restored.

For this reason susceptibility to the polarizational errors of the raised on height/altitude antenna systems of radio direction finder to conveniently characterize by the relation of effective height for the horizontal and vertical electric fields H_2/H_1 for any angle of incidence in the determined parameters of soil.

This same the parameter is convenient for the characteristic of susceptibility to the polarizational errors of phase radio direction finder.

To compare in the relation to the polarizational errors of two radio direction finders, if the characteristic of susceptibility to polarizational errors of one of them is assigned in the form H_2/H_1, and another in the form $\Delta\theta_{cr}$, is difficult.

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At the large angles of the slope of a front from the ionosphere of the radio waves reflected of radio direction finders with the spaced vertical wire antennas, increases the relation of voltages from horizontal and vertical components electric field and increase polarizational errors. Occurs this because with an increase β reception of vertical field, proportional $\cos^2 \beta$, decreases, but the reception of horizontal field barely depends on angle β. On short waves the range of the steep waves when increase polarizational errors, is located on distances to 400-500 km of radio transmitter.

Furthermore, with an increase in the wavelength deteriorates the relationship/ratio of vertical and horizontal receptions, since effective height for a vertical field falls, and for horizontal it is retained approximately constant.

System with the diverse framework does not possess these deficiency/lacks, since in it voltages from the reception of vertical and horizontal fields are proportional to $\cos \beta$, i.e., they fall simultaneously with an increase β. The relation of the reception of

horizontal and vertical fields of this system does not depend the odes of wavelength.

Let us give some statistical materials, characterizing RDF systems with spaced antennas [1.10]

Fig. 6.11 and 6.12 gives percentage error distributions of bearings in night time on systems by U-shaped shielded, balanced with transformers and obtained on rotatable loop. These figures are the result of processing several thousands of night observations on medium-frequency waves.

It is interesting to compare the mean square errors of the indicated systems, obtained as a result of these experiments. In Fig. 6.11 mean square errors are equal: U-shaped system shielded 2.4°, rotatable loop 12.4°. In Fig. 6.12 mean square errors are equal: system balanced with transformers 1°, rotatable loop 8.4°.

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If all results are converted to the mean square error of rotatable loop into 12.4°, then we will obtain the following comparative errors of systems in the degrees: rotatable loop 12.4; system U-shaped shielded 2.4; transformer balanced (balanced) system 1.5

From the comparison of given data with the error of standard wave of framework (35.3°) it is evident that the average quadratic operating error comprises approximately 35-40o/o of error of standard wave. Hence it is possible to draw a conclusion about the permissible error of standard wave with the assigned operating error.

Were the attempts to use for the compensation for the reception of horizontal field to the framework the supplementary horizontal wires, connected with framework [6.6]. However, such systems did not find practical application/use as a result of the high dependence of the required compensation on frequency and soil of the installation of radio direction finder.

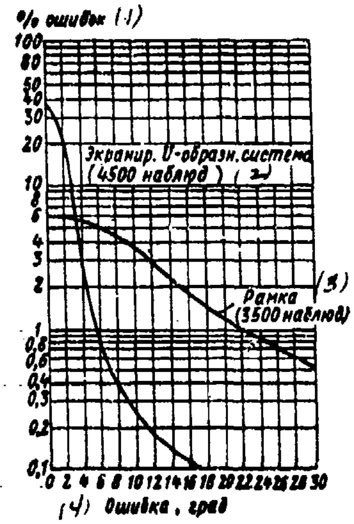

Fig. 6.11.

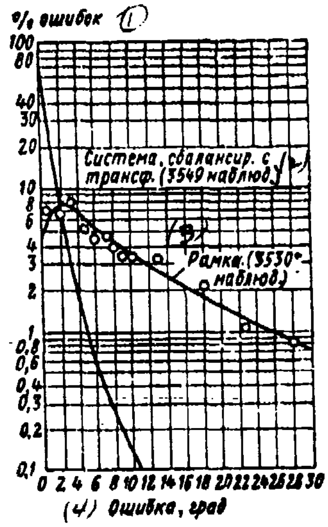

Fig. 6.12

Fig. 6.11. Comparison of the accuracy of direction finilinj by the framework and tha U-shaped system.

Key: (1) Errors. (2). Shielded U-shaped system (4500 obsarvations). (3). Framework (3500 observations). (4). Error, deg.

Fig. 6.12. Comparison of the accuracy of direction finding by the framework and the balanced system with transformers.

Key: (1) errors. (2). System, balanced with transformer (3549 observations). (3). Framework (3530 observations). (4). Error, deg.

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Circular antenna systems with acute/sharp directional characteristic to a lesser degree are subjected to polarizational errors, than system with cosinusoidal characteristic, for the following reasons:

a) the average value of the projection of feeders on any direction smaller than the diameter of spacing of the antennas; consequently, the less effective height for a horizontal electric field;

b) the voltages in feeders store/add up not cophasally;

c) feeders have the best screening because of the large size/dimensions of wire gauze of grounding.

The protection of phase systems on polarizational errors depends on the workmanship of an antenna-feeder system.

System with pulse transmission.

To decrease the polarizational errors is possible, applying in the oriented transmitter pulsing. For the realization of this method of direction finding transmitter, intended for direction-finding, must be supplied with special equipment/device for the emission/radiation of momentum/impulse/pulses. In this case not the output of receptor is included the cathode-ray tube on screen of which it is observed both momentum/impulse/pulse of terrestrial wave and the momentum/impulse/pulses, reflected from ionospheric layers. Virtually it is possible to obtain several (to 6 and more) reflections from layer E, several (two and more) reflections from layer F, and also the diffuse reflections, near and remote. Images on the screen of cathode-ray tube lag behind each other to the time, caused by the supplementary distance of propagation, determined by the height/altitude of reflecting layer and by level of reflection.

The schematic diagram of equipment/device for a pulse radio direction finder is shown in Fig. 6.13.

The momentum/impulse/pulses, taken by the framework or another directional antenna, create voltage on one pair of plates of cathode-ray tube. To another pair of plates, is fed the sweep voltage, synchronized by the local oscillator of low frequency whose frequency is regulated to its coincidence with the pulse repetition frequency of transmitter. Then on the screen of cathode-ray tube are

obtained on the medium-frequency waves of image, analogous Fig. 6.14.

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The first peak corresponds to terrestrial wave (Fig. 6.14a), the second (but sometimes and the third), that came is somewhat later, corresponds to reflection from layer E (Fig. 6.14b), E_1 - the first reflection from layer E, E_2 - the second so forth.

Direction finding is reduced to the rotation of framework or search coil of goniometer and finding of the situation, when disappears the image of the first peak from ground ray (Fig. 6.14c). This method of direction finding is led to release from polarizational errors, but it it is possible to use with mission time of momentum/impulse/pulse only up to determined distances from transmitter, thus far the pulse duration is not greater than difference in the transit time of terrestrial and reflected waves. Knowing the pulse duration and being given the height/altitude of reflecting layer, it is possible to calculate the maximum distances of direction finding.

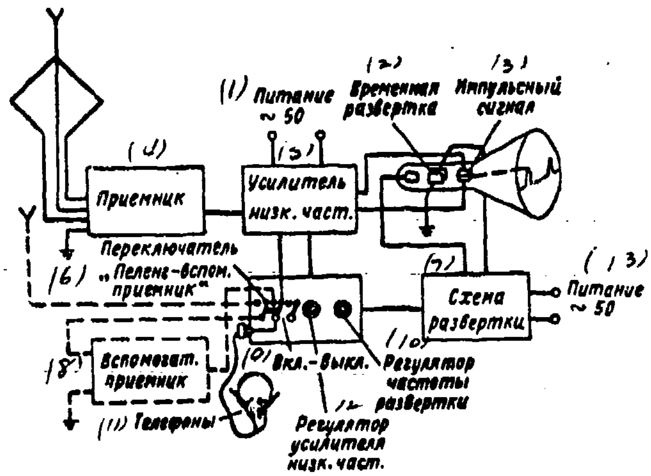

Fig. 6.13. Schematic diagram of pulse radio direction finder.

Key: (1). Feed. (2). Time/temporary scan. (3). Pulse signal. (4). Receiver. (5). Amplifier is low. it is frequent. (6). Switch "Bearing-Aux. receiver". (7). Circuit of scan. (8). Aux. receiver. (9). inclusive-off. (10). Frequency regulator of scanning/sweep. (11). Telephones. (12). The regulator of l-f amplifier. (13). Feed.

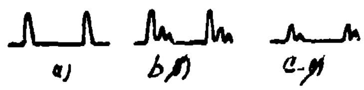

Fig. 6.14. Image on screen of cathode-ray tube. (Sweep frequency is two times greater than the pulse repetition frequency).

Fig. 6.15. Images of momentum/impulse/pulses under different conditions of propagation of short radio waves.

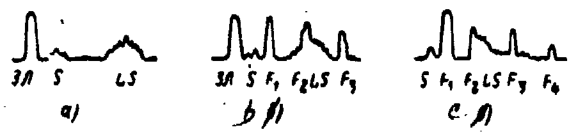

Fig. 6.15.

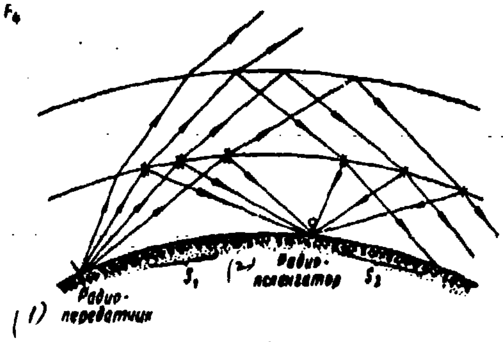

Fig. 6.16.

Fig. 6.16. Ranges of diffuse reflection.

Key: (1). Radio transmitter. (2). Radio direction finder.

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On short waves at tube face, appear the images of ray/beams as at Fig. 6.15.

In Fig. 6.15a after the image of ground ray (ZL) there are two groups of scattered radiation S and LS whose sources are at different distances from direction finder.

Figures 6.16 depicts the predicted paths of radiowave propagation for obtaining image analogous Fig. 6.15a. By crosses are

marked the possible ranges of diffuse reflection from the ionosphere (S_1 - the near range of scattering, S_2 - the remote range of scattering). In Fig. 6.15b the detachable wave follow the reflections from layer F (single F_1, twofold F_2 and triple F_3). Image F_2 is lost against the background of the scattered (distant) reflection LS. In Fig. 6.15c terrestrial wave is absent. Image F_2 almost overlaps with diffuse reflections. Under specific conditions can appear also the images of reflections from layer E.

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It is possible to carry out direction finding on any of the depicted momentum/impulse/pulses. On terrestrial wave on short waves, the bearing is reliable, thus like the error due to reflection are absent. Bearing on first reflection from layer E is most reliable of the bearings on the reflected beams, since from layer E wave comes most hollow, also, with minimum lateral deviations.

A deficiency/lack in the method is the wide emission band of transmitter and the creation because of this of the interferences in pulse transmission to adjacent channels of communication/connection. The receiver of radio direction finder must possess the broad passband of frequencies. So, with the duration of momentum/impulse/pulse 300 μs the passband of the frequencies of the

receiver must be order 5 kHz, with the duration of momentum/impulse/pulse 0.5 μs, the passband of the fraquencies of the receiver must be $3 \cdot 10^6$ Hz.

Another method of pulse direction finding is described into § 8.7.

6.4. Lateral deviations of the radio waves of skip band.

With the reflection of electromagnetic waves from the ionosphere on short waves besides a change in the plane of polarization, are observed the phenomena of lateral deviations and radio interference.

As a result of the fact that at reflecting layer has the changing in time horizontal gradient of ionization, the surface of layer seemingly become undulating. As a result in any point of layer, appear the changing in time slope/inclinations, calling the lateral deviations of radio waves, i.e., the bearing error and change in the angle β. Table 6.1 gives the root mean square values of the angles of the slope of layers E and F and the average speed of their change into the daytime are frequent under conditions of calm ionosphere [6.11]. The RMS slope angles change approximately inversely

proportional to frequency. Rate of change in the slope/inclination lie/rests within limits 0.4-0.5 deg/min. In night hours in summer the angles of the slope of layer increase not considerably in comparison with the daytime, while in night hours in winter and during the ionospheric disturbances - to two and are more once.

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When wave passes in the plane, which separate/liberates the darkened part of the earth/ground of that which was illuminated is obtained the horizontal gradient of ionization, which appears as a result of illumination change. Calculations show that for this reason the slope/inclination of layer cannot be more than 1°.

The period of slope deviation for any direction of reflection varies from one to 30 and more minutes.

Any slope/inclination of layer can be decomposed on transverse and longitudinal. The lateral deviations of radio waves are caused by the transverse slope/inclinations when standard to layer is deflected in the plane, perpendicular to the plane of great circle, the containing direction of propagation. Longitudinal slope/inclinations (standard to layer is deflected in the plane of great circle) are produced change in the angle β. The oscillation/vibrations of the

slope/inclinations of layer are led to slow changes in the bearing in time with the period of change from 1 to 30 min and more.

Let us designate: ψ is an angle of the rotational axis of layer with the plane of wave front; δ is angle of the slope of layer of relatively horizontal plane; $h_{сл}$ is a height/altitude of layer; D - distance is transmitter - direction finder.

Table 6.1. Effective values of the slope/inclination of layers according to data [6.11].

(1) Отражение от слоя	(2) f. Мгц	(3) Эквивалентная высота, км	(4) Средний квадратический угол наклона, град			(5) Средняя скорость изменения угла наклона град/мин	
			NS	EW	(6) общий	NS	EW
E	3,7—7,3	100—130	1,1	1,4	1,8	0,5	0,5
F—1 (7) (обыкн.)	4,3—4,6	230—330	1,4	1,4	2	0,6	0,5
F—2 (7) (обыкн.)	5,0—7,7	260—520	1,8	1,4	2,4	0,4	0,3
F—2 (8) (необыкн.)	5,0—7,7	240—500	1,5	1,1	1,8	0,4	0,4

Key: (1). Reflection from layer. (2). MHz. (3). Equivalent height/altitude, km. (4). Average quadratic angle of the slope, deg. (5). Average rate of change of angle of the slope deg/min. (6) common/general/total. (7) (ordinary). (8) (extraordinary).

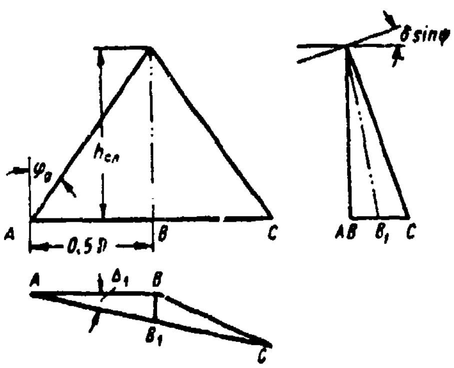

Fig. 6.17. Error due to lateral deviation.

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Then with single reflection, as this follows from Fig. 6.17 (earth/ground it is assumed to be the plane), the azimuth deviation of radio waves Δ_1 from the plane of great circle transmitter - direction finder will be

$$\operatorname{tg}\Delta_1=\frac{2h_{cs}}{D}\operatorname{tg}(\delta\sin\psi). \qquad (6.3)$$

Than higher layer and the lesser the distance between the transmitter and the direction finder, the greater the error Δ_1 due to the slope/inclination of layer.

In (6.3) $\frac{2h_{cs}}{D}=\operatorname{ctg}\varphi_0$, where φ_0 is an angle of incidence in the wave on layer; therefore

$$\operatorname{tg}\Delta_1=\operatorname{tg}(\delta\sin\psi)\operatorname{ctg}\varphi_0. \qquad (6.3')$$

Upon consideration of the sphericity of the earth/ground formula (6.3') assumes the form

$$\operatorname{tg}\Delta_1=\operatorname{tg}(\delta\sin\psi)\operatorname{ctg}\varphi_0\cos\Phi,$$

where Φ is a half of the central angle between the transmitter and the direction finder. For the small distances

$$\cos\Phi\approx 1 \text{ и } \operatorname{tg}\Delta_1=\operatorname{tg}(\delta\sin\psi)\operatorname{ctg}\varphi_0.$$

For the large distances

$$\cos\Phi\approx\sin\varphi_0 \text{ and } \operatorname{tg}\Delta_1=\operatorname{tg}(\delta\sin\psi)\cos\varphi_0.$$

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If one considers that with reflection from layer F_2 the angles of incidence lie/rest within limits φ_0 = 30-70°, then must be fulfilled when ψ = 90° equality tg Δ_1 = (0.34-1.73) tg δ or approximately at small angles δ

$$\Delta_1 = (0,3 \div 1,73)\,\delta.$$

I

In the presence of two reflections, if the first of them is characterized by values δ_1, ψ_1, $h_{\text{сл}1}$, the second - by values δ_2, ψ_2, $h_{\text{сл}2}$, the complete lateral deviation of emission/radiation will be

$$\operatorname{tg}\Delta_2 \approx \frac{2h_{\text{сл}1}}{D}\left(\operatorname{tg}\delta_1 \sin\psi_1 + 3\frac{h_{\text{сл}2}}{h_{\text{сл}1}}\operatorname{tg}\delta_2 \sin\psi_2\right). \quad (6.4)$$

A

Assuming the statistical characteristics of lateral deviations at two mirror points identical and $h_{\text{сл}1} = h_{\text{сл}2} = h_{\text{сл}}$, we will obtain that depending on the statistical communication/connection between slope/inclinations δ_1 and δ_2 the variance of error of bearings will be: - in the absence of the correlation between δ_1 and δ_2 $\overline{\Delta_2^2} = 10\overline{\Delta_1^2}$,

- during the total correlation $\overline{\Delta_2^2} = 16\overline{\Delta_1^2}$.

With an increase in the number of reflections of error due to lateral deviations rapidly they increase. In all cases with the

number of reflections, greater than one, the nearer the mirror point to receiver, the more powerful affects the reflection at this point the error of lateral deviation.

Figures 6.18 gives for a radio direction finder with cosinusoidal directional characteristic the averaged dependence on distance of variance of error due to lateral deviations separately for day and night [11.4].

It is experimentally establish/installed that if we conduct direction finding at two points, equally removed from radio transmitter and which are located at a distance to 80-100 km one from another, then the lateral deviations of bearings at these points approximately coincide, i.e., the coefficient of the correlation of the slow oscillations of bearings is close to unity.

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Thus, the conditions of the reflection of radio waves from the ionosphere at the points, spread on 40-50 km, are approximately identical. With an increase in the distance between the reflecting points, the coefficient of the correlation of slow oscillations falls.

With the distance of mirror points approximately by 100 km (between radio direction finders distance approximately 200 km) the correlation coefficient between slow oscillations is already close to 0, coincide well only statistical characteristics of the oscillations.

Simultaneously with the oscillations of bearings with period 1-30 min are observed changes in the bearings with much more prolonged period. The nature of these oscillations is not completely establish/installed. Is observed the difference of average diurnal bearings of relatively average bearings for long time. There are seasonal variations of average bearings. These changes in the bearings are not explained by the effect of near environment, since the introduction of corrections on the local oscillator does not eliminate them.

Reason, apparently, consists in change in time of the characteristics of soil and parameters of the environment with the antenna of system.

Since so on multiple-pronged radiowave propagation smallest lateral deviation has the wave, which underwent one reflection, was proposed the method of a decrease in the lateral deviations with the reception of telegraph signals by the isolation/liberation of wave

with one reflection.

Figures 6.19 depicts the echo telegraph signal in the presence of two ray/beams.

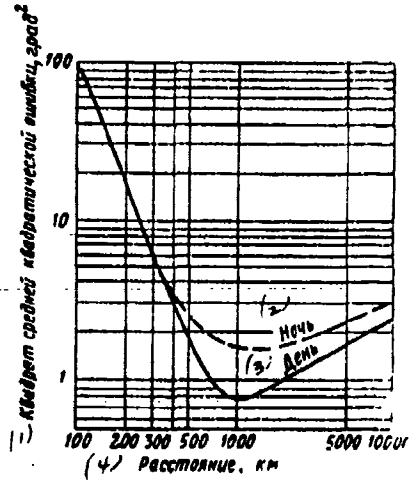

Fig. 6.18. Dependence of errors due to lateral deviation from distance.

Key: (1). Square of mean square error, deg². (2). Night. (3). Day. (4). Distance, km.

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For convenience in the image, component signals are shown one under another. The resulting signal can be equal or to sum (solid line), or difference in the components (dotted line). In radio direction finder it is possible to have such equipment/device which would open/disclose receiver from each beginning of telegraph signal to the time, corresponding to presence one of first reflection. Then all

other echo signals are excluded and bearing must have the smallest error.

For this, realization they utilize following. Between the first and second reflections from layer F, must be the differense in time, required for the passage of path, of approximately equal to two height/altitudes reflecting layer. Considering that height/altitude of layer 300 km, the time between first and following reflections must be on the order of 2 ms. Into radio direction finder is inserted special equipment/device with the aid of which they open/disclose its only to 2 ms from the beginning of each telegraph sign, and thus is realized direction finding on the first reflection. In two-channel automatic direction finder without the acceptance of spesial measures on tube face, are obtained separately the bearings, which correspond to the first reflection and cumulative effect of all reflections.

For the larger duration of premise/impulses, image brightness of the first reflection is much less than total signal, which impedes reading on the first reflection.

Usually for the direction finding of radio stations, it is abstract/removed time less than it is necessary for the averaging of the slow oscillations of bearings. Therefore, it is not represented to possibility of decreasing the error due to the slow oscillations of bearings by means of the averaging of readings.

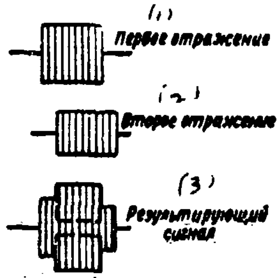

Fig. 6.19. Addition of the first and second reflections of telegraph signal.

Key: (1). First reflection. (2). Second reflection. (3). Resulting signal.

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A decrease in the lateral deviations it is not possible to also achieve by the simultaneous direction finding one and the same of radio station on two diverse radio direction finders and by the calculation of average bearing, since the coefficient of the correlations of lateral deviations at two points at a distance to 100 km is close to unity.

In radio direction finders with the larger separation of antennas during the multiple-pronged propagation of error due to

lateral deviations, is less than in cosinusoidal systems, because of three-dimensional/space selectivity and the preferred reception of more intense wave with one reflection, when lateral deviations have smaller values.

6.5. Radio interference of skip band.

Radio waves can simultaneously be reflected from layers E, F_1, F_2, after undergoing reflection one, two and more once. With reflection from any layer due to the local heterogeneities of layer ("roughness") occurs the partial scattering of radio waves. Furthermore, with reflection is observed magnetoionic splitting/fission into usual and extraordinary waves. It is concealed by shape, the oriented signal can consist of the large number of component ray/beams, which interfere between themselves, moreover each of the reflected beams is usually accompanied by the beam (cone) of scattered waves.

In the examination of the passage of the wave through the ionized layer, it is assumed that the ionization of layer in vertical and horizontal directions changes independently according to normal law, moreover reflection affects the local heterogeneity of layer in

range with the size/dimensions of sides approximately a = 500 m.

On output from layer, the power of the incident wave (after losses in layer) is distributed between the correctly reflected beam and the beam of scattered waves. The relation of the power of fundamental ray/beam and the total power of the beam of scattered waves [6.15]

$$q=\sqrt{\frac{e^{-\Phi_{\text{д}}^2}}{1-e^{-\Phi_{\text{д}}^2}}},$$

where $\Phi_{\text{д}}^2$ — the dispersion of equivalent phase displacement as a result of the passage through the layer, which depends on the characteristics of layer and frequency.

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At low values $\Phi_{\text{д}}$ < 0.5 rad

$$q=\frac{1}{\Phi_{\text{д}}}.$$

Experiment showed that at frequency 5 MHz q = 2.5 and $\Phi_{\text{д}}$ = 0.3 rad.

The angular distribution of power in the beam of scattered waves has the dispersion

$$\theta_0^2=\frac{\lambda^2\Phi_{\text{д}}^2}{2\pi^2a^2\left(1-e^{-\Phi_{\text{д}}^2}\right)}.$$

with small $\theta_0=\frac{\lambda}{\sqrt{2}\pi a}$

and power in beam has the normal law of distribution.

The coefficient of the correlation of the angular spectrum in two antennas at a distance 2b is designed from the formula

$$R(2b)=\frac{e^{-\Phi_{\text{д}}^2}}{1-e^{-\Phi_{\text{д}}^2}}\left[e^{\Phi_{\text{д}}^2 e^{-\left(\frac{2b}{a}\right)^2}}-1\right].$$

When $\Phi_{\text{д}}$, smaller than 0.5 rad,

$$R(2b)=e^{-\left(\frac{2b}{a}\right)^2}.$$

The root mean square value of a supplementary phase difference at these two points is determined from $\Phi_{\text{д}}$ and R(2b):

$$v_0=\Phi_{\text{д}}\sqrt{1-R(2b)}=\frac{1}{q}\sqrt{1-R(2b)}$$

and the root-mean-square deviation of bearing σ_0, caused by scattering ray/beam with the reflection which is determined from condition $4\pi b/\lambda \sin\sigma_0=v_0$, will be

$$\sigma_0\approx\frac{\lambda v_0}{4\pi b}=\frac{\lambda\sqrt{1-R(2b)}}{4\pi bq}\approx\frac{\lambda\Phi_{\text{д}}}{2\sqrt{2}\pi b}\sqrt{1-e^{-\left(\frac{2b}{a}\right)^2}}.$$

With small 2b/a

$$\sigma_0=\frac{\lambda\Phi_{\text{д}}}{\sqrt{2}\pi a}.$$

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Calculations show that scattering radio waves is determined mainly by layer E, even when reflection originates from layer F.

Since to direction finder can come several ray/beams, which underwent single and multiple reflections, due to the interference of these ray/beams are observed the oscillations of bearings within large limits. For example, when the radio direction finder are approached simultaneously two ray/beams of approximately identical intensity and the angle between them is very small, but a phase difference is close to 180°, is obtained the deviation of bearing to ± 90°. In Fig. 6.20 OA - the fundamental reflected beam, which is propagated along arc of the great circle; OB is the second supplementary ray/beam, which forms small angle δ with OA. The strength of the magnetic field of the second ray/beam H_2 has a phase difference relative to the magnetic intensity of the fundamental ray/beam H_1, close to 180°. The resulting magnetic field $H_{рез}$ corresponds to direction of propagation OC. This direction is displaced with respect to the fundamental ray/beam OA in angle ψ, close to 90°.

A phase difference between the interfering ray/beams can take any values, and the displacement of the resulting direction will vary

within the range of 0 to ± 90°.

The height/altitude and the characteristics of ionosphere layers, which affects reflection and scattering radio wave, rapidly change with period from fractions of a second to five and more seconds. The period of rapid oscillations increases with a decrease in the frequency and in night hours.

With the same period occurs a change in amplitude and phase of component waves, and also of amplitude and direction of the resulting field. Simultaneously changes its polarization. Appear the rapid oscillations of bearings due to the phenomenon of interference and alternating/variable polarization of the field reflected, which reach to ± 90°.

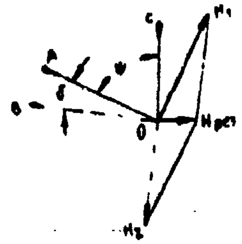

Fig. 6.20. Onset of large error due to two-beam interference.

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In the presence of radio interference, change the size/dimensions and the form of the image of bearings on the screen of the cathode-ray tube of two-channel radio direction finder - from linear the bearing becomes elliptical and even circular.

Let us examine, as they change in the presence of two-beam interference bearing error, the length of image, i.e., the length of the transverse, and the ratio of small and transverses of image on cathode-ray tube.

In accordance with formulas (5.11) and (5.10) for a radio direction finder with cosinusoidal directional characteristic for a bearing error Δ we have the expression

$$\operatorname{tg} 2\Delta = \frac{k^2 \sin^2 2\psi_0 + 2k \sin \psi_0 \cos \varphi}{1 + k^2 \cos 2\psi_0 + 2k \cos \psi_0 \cos \varphi};$$

the length of the transverse of image it will be

$$B = \sqrt{\cos^2\Delta + k^2 \cos^2(\psi_0 + \Delta) + 2k \cos \Delta \cos (\psi_0 + \Delta) \cos \varphi};$$

the ratio of the axes of ellipse is determined from the formula

$$\frac{A}{B} = \frac{\sqrt{\sin^2 \Delta + k^2 \sin^2 (\psi_0 + \Delta) + 2k \sin \Delta \sin (\psi_0 + \Delta) \cos \varphi}}{\sqrt{\cos^2\Delta + k^2 \cos^2 (\psi_0 + \Delta) + 2k \cos \Delta \cos (\psi_0 + \Delta) \cos \varphi}}.$$

For a special case of two-beam interference at the angle between them $\psi_0 = 10°$ and in the relation of strength for ray/beams k = 0.9 are designed the dependences of the bearing error Δ, of the length of the transverse of image B and the ratio of axes A/B on a phase difference φ between the strengths of the field of ray/beams.

Results are represented in Fig. 6.21a.

From the curves of Fig. 6.21a, it is evident, that at the maximum length of the transverse of image (B = 1) and the minimum value of ratio A/B (it corresponds to a phase difference of the interfering ray/beams φ = 0) bearing error $\Delta \approx \frac{\psi_0}{2}$. Error and the ratio of the axes of ellipse are little affected until the length of image becomes less than 25-30o/o of the maximum and A/B > 30-40o/o.

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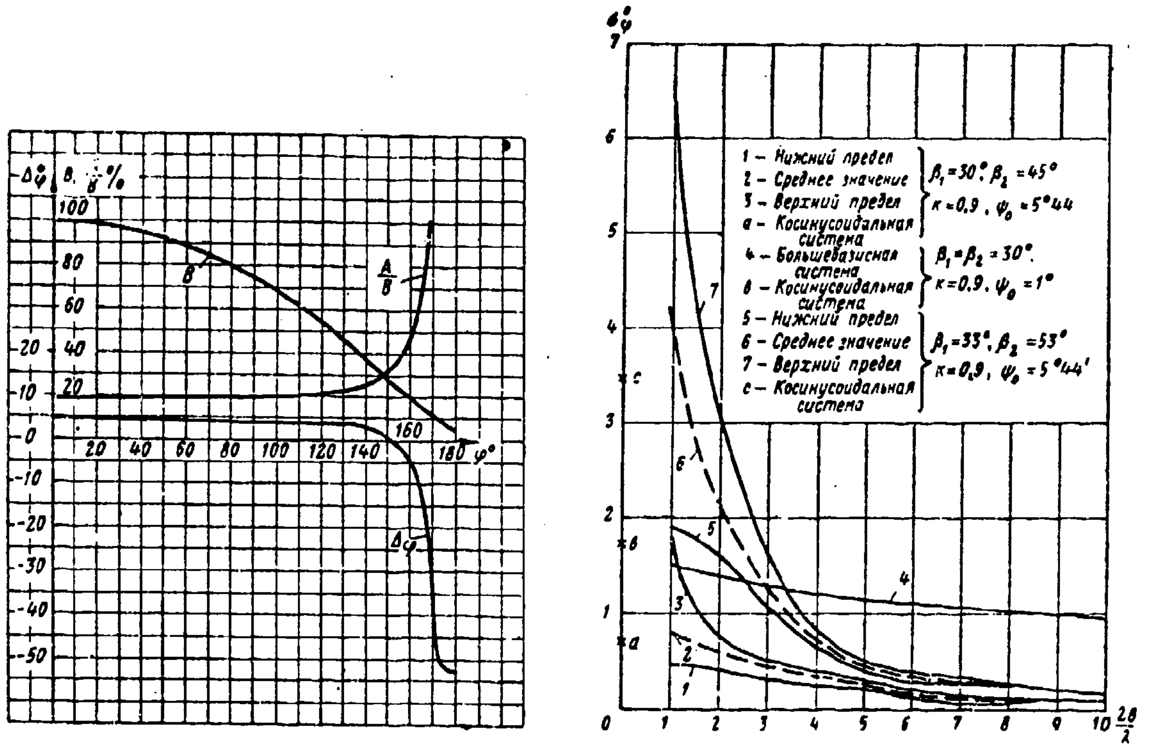

Fig. 6.21. The interference errors of radio direction finder with two ray/beam: a - dependence of B, A/B, Δ on a phase difference with cosinusoidal diagram (k = 0.9; ψ₀ = 10°); b - the dependence Δψ on the separation of antennas for a system with the cyclic measurement of phase.

Key: (1). Lower limit. (2). Average value. (3). The upper limit. (a). Cosinusoidal system. (4). Large-base system. (b). Cosinusoidal system. (5). Lower limit. (6). Average value. (7). The upper limit. (c). Cosinusoidal system.

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With even smaller size/dimensions of the major axis (and the large ratios A/B) the error can grow considerably. Therefore, if operator takes single reading with small image size and large ellipticity, he can obtain large error.

In Table 6.2 are designed the bearing errors, calculated along the transverse of image on the cathode-ray tube of two-channel radio direction finder in the presence of two-beam interference.

The rapid oscillations of bearing due to interference can be averaged for time of direction finding by means of the taking of several readings and calculation of average bearing. Since all the values of a phase difference φ from 0 to 2π are equiprobable, if we take the particular readings through equal time intervals, without considering change in the signal amplitude and the ellipticity of image, the error of the averaged bearing it is determined by its arithmetic mean value without taking into account of sign Δ_{cp}:

$$\Delta_{cp} = \bar{\Delta} = \frac{1}{2\pi}\int_0^{2\pi}\left[\frac{1}{2}\operatorname{arc\,tg}\frac{k^2\sin^2 2\psi_0 + 2k\sin\psi_0\cos\varphi}{1 + k^2\cos 2\psi_0 + 2k\cos\psi_0\cos\varphi}\right]d\varphi.$$

Integral determination in the general case difficultly. At the low values of the errors when $\Delta \approx k \sin \psi_0 \cos \varphi$, integration gives zero value for arithmetic mean error, i.e.,

$$\Delta_{cp} = \frac{1}{2\pi}\int_0^{2\pi} k\sin\psi_0\cos\varphi = 0.$$

Table 6.2. Bearing errors with $\varphi = 0$ and $\varphi = \pi$ for the different values of k and ψ_0.

k	ψ_0, (1) град	Ошибка в градусах при значениях (2)	
		$\varphi = 0$	$\varphi = \pi$
0,9	1	0,5	−9,4
	3	1,3	−31
	5	2,3	−43,3
	10	4,8	−55,3
0,5	1	0,3	−1
	3	1	−3
	5	1,7	−4
	10	3,3	−10

Key: (1) deg. (2). Error in degrees at values.

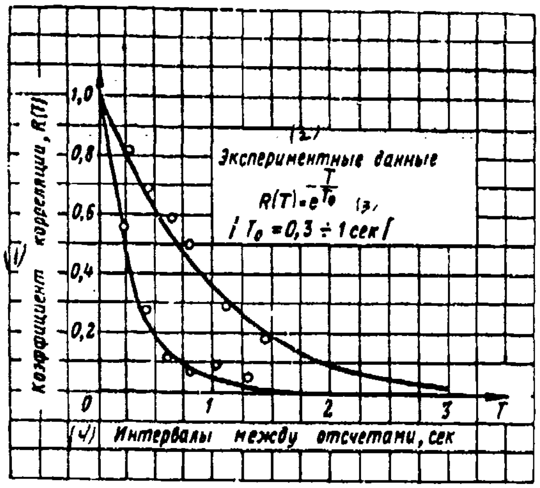

Fig. 6.22. The dependence of the correlation of errors on the time between the observation: ooo - experimental data; —— - calculated curves.

Key: (1). Correlation coefficient R(T). (2). Experimental data. (3) s. (4). Intervals between readings, s.

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Numerical integration error function (Fig. 6.21a) shows that by means of the averaging of bearing it is possible to obtain an error

less than with ϕ = 0. However, if one considers that virtually it is possible to take readings during amplitude reduction of signal only to 4-5 times relatively maximum, then a decrease in the error not very greatly in comparison with reading with ϕ = 0.

Does arise question, as frequently one should take the readings of bearing, as does affect the averaging time for mean error and of which decrease in the rapid oscillations of bearing it is possible to achieve virtually?

Figures 6.22 gives two obtained experimentally the curves of the dependences of the coefficient of the correlation of readings R(T) on time interval T between readings [6.14]. The theoretical dependence R(T) on T takes the form

$$R(T)=e^{-\frac{T}{T_0}},$$

moreover from the curves of Fig. 6.22 it follows that T_0 = 0.3-1 s.

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The curves of Fig. 6.22 show that if the time interval among readings is more than 1-2 s, then the coefficient of the correlation of adjacent readings is close to zero, readings become independent variables.

The coefficient of the correlation of rapid oscillations on the radio direction finders, spread more than on 200-400 m, it is close to zero. Therefore the averaging of rapid oscillations is possible also by the calculation of the average value of the bearings, undertaken simultaneously on several radio direction finders, spread for distance larger 200-400 m.

The effect of averaging time for the dispersion of the rapid oscillations of bearings is represented in Fig. 6.23. Averaging was realized automatically, i.e., readings were taken after every 0.08 s. For averaging time 10 s (125 readings) the dispersion decreases 4 times (decrease in the root-mean-square deviation 2 times), a decrease in the dispersion 10 times is achieved at the averaging of the readings, undertaken approximately after 40 s [8.32].

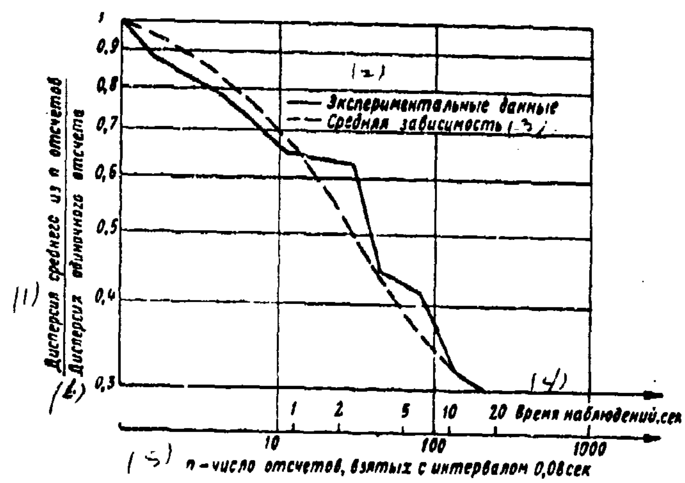

Fig. 6.23. Effect of averaging time for the dispersion of the rapid oscillations of bearings.

Key: (1). Dispersion of average from the p of readings. (2). Experimental baths. (3). Average dependence. (4). Time of observations, s. (5). n - number of the readings, undertaken with interval 0.08 s. (6). Dispersion of single reading.

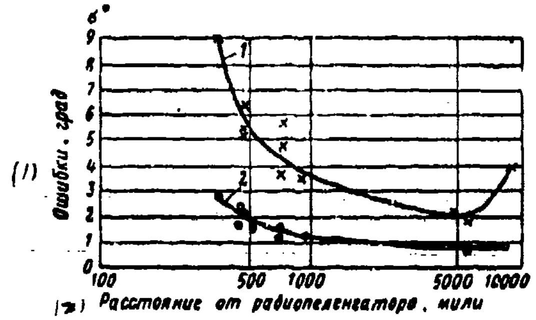

Fig. 6.24. The dependence of the root-mean-square deviation of bearings from the distance: 1 - single readings; 2 - averaged readings; x and o - experimental data.

(1). Errors, deg. (2). Distance from radio direction finder, mile.

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Processing the results of the direction finding of radio stations with ranges 1000-5000 km on goniometric spaced-antenna direction finder showed that the dispersion of single readings in daytime hours has average value of 10 deg^2. It can be decreased 10 times, i.e., it is led to 1 deg^2 by means the averaging of readings during 5 min (10-12 readings at intervals 20-30 s). Dispersion on the order of 1 deg^2 have average for 1-2 hours bearings and the average

diurnal bearings of relatively mean for the wide interval time. Therefore an increase in the averaging time of bearings is greater than that that is required for a decrease in the dispersion 10 times, i.e., to 1 deg², it is inexpedient. At the same time, with taking more frequent than after 20-30 s of the readings of this decrease in the dispersion (10 times) it is possible to achieve with direction finding during 30-40 s (Fig. 6.23).

Thus, under conditions of the normal passage of radio waves in daytime hours it is expedient to average bearing during 30-40 s.

In night hours and in the low-frequency part of the range of the short waves when the period of rapid oscillations increases, time of direction finding is expedient as far as possible to increase.

Figures 6.24 depicts the obtained on goniometric radio direction finder dependence on the distance of the average quadratic oscillations of the single readings of bearing relative to average values for 10 min. In this same figure are given average quadratic oscillations of bearings, automatically averaged after 41 s (16 readings at intervals between readings 2.56 s).

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As a result of the averaging of fluctuation, they decrease from two to four times (on the average 3.3 times). Figures 6.24 shows that the minimum spreads of bearings are observed at a distance 1500-4000 km. With a decrease in the distance, they increase faster than with an increase.

During application/use in radio direction finder by the antenna of system with acute/sharp directional characteristic due to the manifestation of the three-dimensional/space selectivity of the fluctuation of bearings from radio interference have the smaller values than in as an antenna to system with cosinusoidal characteristic.

We give the following comparative results of the separate series of observations on goniometric radio direction finder and on radio direction finder with acute/sharp characteristic direction [6.16]. Dispersion of readings due to the rapid fluctuations of the bearings:

1) layer F_1, frequency 11 MHz, in the daytime:

- on goniometric radio direction finder 1.6 deg^2;

- on radio direction finder with acute/sharp directional characteristic 0.17 deg^2;

2) frequency 5 MHz, at night:

- on goniometric direction finder 18.2 deg^2;

- on radio direction finder with acute/sharp directional characteristic 1.4 deg^2.

With twofold reflection the rapid fluctuations of bearings grow/rise. The value of rapid fluctuations depends on the character of soil at mirror point from the earth/ground. It is experimentally establish/installed that the smaller fluctuations are observed, when after the first reflection from the ionosphere wave falls on sea surface, but not when it falls to the earth. This is explained to the facts that the coefficient of scattering radio waves with reflection from sea surface is less than from the earth/ground.

Let us designate the dispersion of the rapid fluctuations of

bearings with single reflection from the ionosphere by $\overline{\Delta^2_{\delta 1}}$. The dispersion of the rapid fluctuations of bearings after the second reflection from the ionosphere, if after the first wave reflection falls on sea surface and angular scattering with reflection from sea it does not increase, will be

$$\overline{\Delta^2_{\delta 2}} = 2\frac{\cos^2\delta_1}{\cos^2\delta_2}\overline{\Delta^2_{\delta 1}},$$

where δ_1 and δ_2 are angles of incidence in the wave on ionospheric layer with the first and second reflections.

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Factor $\cos^2\delta_1/\cos^2\delta_2$ considers that the bearing error is proportional sec δ.

If after the first reflection from layer wave falls to the earth, then, by considering the coefficients of reflection from the ionosphere and the earth/ground approximately identical, we will obtain

$$\overline{\Delta^2_{\delta 2}} = 3\frac{\cos^2\delta_1}{\cos^2\delta_2}\overline{\Delta^2_{\delta 1}}.$$

At frequencies, greater than MPCh [- maximum usable frequency], when the ionization of layer is insufficient for the normal reflection of radio waves, can occur reflections from the

individual moved in layer heterogeneities with large ionization or from areas of the ionosphere from the required for reflection given frequency by ionization, that are found aside from other great circle "transmitter - direction finder". In such cases are observed large alternating/variable errors due to large lateral divergences.

Identical large errors with the same radio station can be observed on several radio direction finders, spread up to large distance (to dozens kilometers).

Susceptibility by the antenna of the system of radio direction finder to interference errors is conveniently characterized by the mean square error σ_φ, which is obtained with let us accept two coherent plane electromagnetic waves (main and supplementary), when a phase difference Φ of the strengths of the field of these waves in center with the antenna of the system of radio direction finder varies from 0 to 2π.

It is accepted that they remain constants:

ψ_0 - an angle in the horizontal plane between the directions of the arrival of two waves;

β_1 - angle in the vertical plane of the direction of main wave;

β_2 - angle in the vertical plane of the direction of supplementary wave;

k - relation of the strengths of the field of supplementary and main waves ($k<1$).

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If the error which is obtained relative to the direction of the arrival of main wave under the indicated conditions and with certain phase difference ϕ, to designate by Δ, then

$$\sigma_\varphi^2=\frac{1}{2\pi}\int_0^{2\pi}\Delta^2 d\varphi.$$

We give the formulas of calculation σ_φ for a cosinusoidal system and systems with the cyclic measurement of phase in high frequency.

Cosinusoidal system.

Error Δ is determined by the formula (5.11), in which let us

replace k with relation $\frac{k\cos\beta_2}{\cos\beta_1}=r$ and $p-\psi=\psi_0$. We assume that in view of smallness Δ it is possible to count

$$\Delta^2=\frac{1}{4}\sin^2 2\Delta=\frac{1}{4}\,\frac{\mathrm{tg}^2 2\Delta}{1+\mathrm{tg}^2 2\Delta}.$$

Then

$$\sigma_\varphi^2=\frac{1}{8\pi}\int_0^{2\pi}\sin^2 2\Delta d\varphi=\frac{1}{8\pi}\int_0^{2\pi}\frac{(2r\sin\psi_0\cos\varphi+r^2\sin 2\psi_0)^2}{(1+2r\cos\psi_0\cos\varphi+r^2\cos 2\psi_0)^2+(2r\sin\psi_0\cos\varphi+r^2\sin 2\psi_0)^2}\,d\varphi.$$

At low value ψ_0

$$\sigma_\varphi^2=\frac{1}{8\pi}\int_0^{2\pi}\frac{(2r\psi_0)^2(r+\cos\varphi)^2}{(1+r^2+2r\cos\varphi)^2}\,d\varphi.$$

Producing the replacement of the variables tg(φ/2) = x, after integration we will obtain for σ_φ $\sigma_\varphi^2=\frac{(r\psi_0)^2}{2(1-r^2)}$ (for $\psi_0\leqslant 6^\circ$ the error of formula less than 6o/o).

For any values ψ_0 the expression for σ_φ will be [6.13].

$$\sigma_\varphi^2\approx\frac{r^2\sin^2\psi_0(1-r^2\cos 2\psi_0)}{2(1-r^2\cos 2\psi_0+r^4)}. \qquad (6.5)$$

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Radio direction finder with the cyclic measurement of phase in high frequency.

In §5.3 it is obtained expression (5.25) for a bearing error with the direction finding of two ray/beams:

$$\operatorname{tg}\Delta = \frac{s\cos\chi}{\delta_1 - s\sin\chi}.$$

After the substitution of value for χ from (5.24) we will obtain

$$\operatorname{tg}\Delta = \frac{2s\delta_2\sin\psi_0}{a\delta_1 - s(\delta_1 - \delta_2\cos\psi_0)}, \qquad (6.6)$$

where

$$\delta_1 = \frac{2\pi}{\lambda} b\cos\beta_1;$$
$$\delta_2 = \frac{2\pi}{\lambda} b\cos\beta_2;$$
$$a = \sqrt{\delta_1^2 + \delta_2^2 - 2\delta_1\delta_2\cos\psi_0};$$
$$s = 2\left[\, k\cos\varphi J_1(a) - \frac{k^2}{2}\cos\varphi J_1(2a) + \dots\right].$$

From expression (6.6) are derived the approximation formulas, which give upper and lower limits for σ_φ [6.13], so that

$$\sigma_{\text{в}} > \sigma_{\varphi} > \sigma_{\text{н}}.$$

The upper limit for σ_{φ}

$$\sigma_{\text{в}} = \frac{\sqrt{2}\,\delta_2 \sin\psi_0 \frac{T}{a}}{\delta_1 - \frac{s}{a}\,|\,(\delta_1 - \delta_2 \cos\psi_0)\,|}. \qquad (6.7')$$

Lower limit for σ_{φ}

$$\sigma_{\text{н}} = \frac{\sqrt{2}\,\delta_2 \sin\psi_0 \frac{T}{a}}{\sqrt{\left\{\left[\delta_1 + \frac{s}{a}\,|\,(\delta_1 - \delta_2 \cos\psi_0)\,|\right]^2 + \left(\frac{s}{a}\,\delta_2 \sin\psi_0\right)^2\right\}}}, \qquad (6.7'')$$

where

$$T^2 = \sum_{m=1}^{p} \frac{k^{2m}}{m^2}\,[J_1(ma)]^2 + R_p,$$

moreover p is taken so as to fulfill the inequality

$$R_p < \frac{0{,}089 k^{2(p+1)}}{a(p+1)^2}\,\frac{1}{1-k^2}.$$

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If angles β_1 and β_2 considerably are distinguished and ψ_0 is small, then it is possible to use the approximation formulas for σ_B and σ_H:

$$\sigma_B = \frac{\sqrt{2}\delta_2 \sin\psi_0}{\delta_1}\frac{T}{a},$$

$$\sigma_H = \frac{\sqrt{2}\delta_2 \sin\psi_0}{\delta_1 + \frac{s}{a}\,|(\delta_1 - \delta_0\cos\psi_0)|}\frac{T}{a}.$$

With the large separation of antennas, when $(2\pi/\lambda)b \longrightarrow \infty$, formula (6.6)-(6.7) they are simplified:

$$\sigma_B = \sigma_H = \sigma_\varphi = \frac{\sqrt{2}\delta_2 \sin\psi_0}{\delta_1}\frac{T}{a}. \qquad (6.8)$$

Formula (6.8) it is possible to use with the sufficient accuracy when $2b \gg 2\lambda$.

The investigation of error σ_φ showed that it increases with an increase ψ_0 (to definite limits, it is analogous to the curves of Fig. 5.6). Error also increases with approach/approximation to the equality of angles β_1 and β_2. Figures 6.21b depicts to the dependence of error σ_φ on the separation of antennas for the following cases:

$\psi_0=5°44'$, $k=0{,}9$, $\beta_1=30°$, $\beta_2=45°$;
$\psi_0=1°$, $k=0{,}9$, $\beta_1=30°$, $\beta_2=30°$;
$\psi_0=5°44'$, $k=0{,}9$, $\beta_1=33°$, $\beta_2=53°$.

From curves it is evident that at low values ψ_0 and $\beta_1 \approx \beta_2$ with an increase in the separation of antennas the error σ_φ little decreases. A noticeable decrease in the error can be obtained with the very large separation of antennas $2b \gg (10-20)\lambda$.

With large ψ_0 an increase in the separation up to $(4-5)\lambda$ is led to noticeable decrease σ_φ. A further increase in the separation affects smaller; furthermore, errors themselves become small.

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Virtually for an effective decrease in the interference errors, it is expedient to increase separation to $(4-5)\lambda$.

The conclusions, obtained for a system with the cyclic measurement of phase in high frequency, are approximately valid also for by the ante of the system of phase radio direction finder and radio direction finder with acute/sharp directional characteristic

with the identical separations of antennas of systems.

6.6. Special feature/peculiarities of direction finding on different wave bands and selection by the antenna of the system of radio direction finder.

Direction finding on very long and long waves (frequency is less than 100 kHz).

The characteristic features of propagation of very low frequencies they are insignificant absorption in soil and ability because of diffraction it is comparatively easy to go around the earth/ground. At short distances from transmitter, smaller 300 km, predominates ground wave. The strength of field does not depend on time of days and year. Spatial waves have small relative to suppress waves the strength of field. At these distances polarization normal and in radio direction finders are applied frame antenna systems with rotatable loop or with two motionless framework. Two immobile frameworks not always coaxial of symmetry - they can be assembled so that are contacted their lateral sides. The noncoincidence of the phases of field on the axis of the symmetry of both framework in

these waves little and is not led to bearing errors.

With an increase in the distance all more predominates sky wave above the terrestrial. From the ionosphere the wave reflected can have, especially into night hours, an abnormally-polarized component of electric field.

Of the framework appear polarizational errors.

According to experimental data at frequencies 16-20 kHz, polarizational errors reach values in daytime hours by the summer:

- at distances 250-600 km approximately to 9°,

- at distances 1000 km approximately to 6°,

- at distances 1500 km approximately to 3°.

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In the night are frequent the errors they grow/rise:

at distances 200-400 km to 20°,

- at distances 1000 km to 10° [6.20].

Is possible the application/use of systems with the spaced vertical wire antennas according to U-shaped diagram with goniometer or with the phase reading of bearing [8.26].

Direction finding on medium-frequency waves (frequency 100-1500 kHz).

Absorption in soil on these waves is more than on long, it increasing with a decrease in the ground conductivity and with an increase in the frequency. Terrestrial wave has vertical polarization, but its range decreases with an increase of absorption in soil.

The most characteristic features from the ionosphere of the waves reflected:

1. Negligible intensity in comparison with ground waves in daytime due to powerful absorption in layers E and D, whereupon on shorter waves absorption grow/rises as a result of approach/approximation to the frequency of gyromagnetic resonance.

On the high latitudes in the daytime hours of winter months, sometimes there is sky wave, since losses under these conilitions are small.

2. Consilerable intensity of these waves in dark time of days at distances from dozen kilometers and more, which is explaiaed by powerful decrease in losses in ionosphere due to decrease in electronic concentration and collision frequency.

3. At three iistances where there are sky waves, is observed fluctuation of intensity, which occurs due to interference of several ray/beams. The oscillatory period is from second to several dozen minutes.

4. Polarization of those who were reflected wills elliptical or linear, that changes orientation in the course of time.

In daytime hours on these waves, predominantly it is not reflected from the ionized layers of the atmosphere of electromagnetic waves, but there is one terrestrial normally-polarized wave.

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Therefore is possible error-free direction finding by the radio direction finders, which apply frame antenna systems.

However, sometimes can come in the daytime and from layer E reflected the wave of weak intensity. Due to the effect of the horizontal component of the electric field of the wave reflected appear the daytime fluctuations of bearings. For this reason the average for day bearings by one and the same radio station can somewhat change from day to day.

Figures 6.25 depicts the average values of bearings for day to 4 radio stations, designated a, b, c, d. The fluctuations of average bearings from day to night reach $\pm 1.5-2°$. Separate bearings during day can have large divergences.

In night hours due to the action of the intense waves reflected are observed the night effects (effect of a normally-polarized field). They most powerful are developed during 1-2 hours to rise and sunset and during 1-2 hours later rise and approach. In southern latitudes powerful night effects are observed in the course of entire night.

Night effects are manifested during the use of frame antenna systems in the following phenomena:

1) bearing does not remain constant, but it changes, "goes for a walk";

2) in radio direction finders with the reading of bearing on the minimum, sometimes do not succeed in obtaining the complete disappearance of audibility and changes the compensation for antenna effects; in two-channel radio direction finders sometimes the image of bearing on cathode-ray tube takes the form of ellipse from the changing orientation of major axis and with the alternating/variable relationship/ratio of axes;

3) occur changes in signal strength, the intensity of signal drops to zero at times (fading).

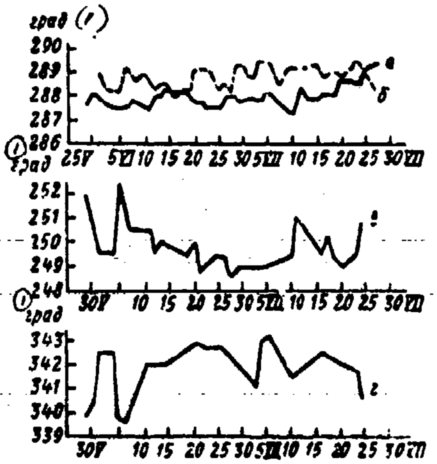

Fig. 6.25. Daytime fluctuations of bearings on medium-frequency waves.

Key: (1). deg.

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More powerful night effects are observed with the direction finding of the radio transmitters whose antennas have large horizontal part (for example, L-shaped), since this antenna emits the horizontally polarized component of electric field. In terrestrial wave this component rapidly attenuates; from the ionized layers of the atmosphere, it is reflected.

On the basis of the available data on prolonged direction finding by the ground-based and ship radio direction finders of different radio stations in the range of waves from 500 to 10000 m to the different hours of days it is possible to arrive at following conclusions [6. 10, 1. 10].

If transmission occurs above sea and there are no errors due to coastal effects, then with loop antennas at distances to 110 km (60 nautical miles) the mean error of direction finding is approximately 2° for the daytime and for night observations. At a distance to 200 km and during the propagation of the waves above sea of 90o/o observations have an error not more than 2°, but maximum error reaches approximately 4°.

During radiowave propagation in dry land, the average diurnal accuracy of direction finding into 2° is obtained at a distance to 40-100 km depending on the wavelength, power of transmitter and character of soil.

With the direction finding of aircraft by the framework from the earth/ground on the generally accepted for this purpose wave λ = 900 m errors due to night effects for distances to 150 km it lie/rests

within the permissible limits.

If on shorter waves, on the order of 400-600 m, the fluctuations of bearings in night hours reach to ±90°, then with the elongation of the wave of the fluctuation of bearings decrease and on wave 10000 m do not exceed 30°. At large distances (more than 2000 km) night errors decrease. The reason for this consists in the fact that the wave reflected comes to direction finder at very small angle to the horizon. In this case,, as can be seen from (6.2), error with the direction finding of an abnormal-polarized field falls.

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In mountainous country the night effects begin to be developed earlier than in plains, which is explained by more rapid weakening of terrestrial wave.

If is placed the problem to carry out direction finding mainly into daytime hours or the 24-hour confident direction finding at small distances (to 150-180 km on sea or to 40-50 km on dry land), then it is possible to use frame radio direction finders.

For the direction finding, free from polarizational errors, are applied predominantly the goniometric systems with the spaced

antennas, assembled according to the diagram of U-shaped system or according to the diagram of transformer system, and also phase radio direction finders.

Direction finding on short waves (1.5-30 MHz).

The conditions of the direction finding of short waves depend on distance of the oriented transmitter, the wavelength and upper-air conditions. For the varied conditions of direction finding, it is expedient to apply the different antenna systems of radio direction finders.

At short distances (to 20-250 km depending on the wavelength, power of transmitter and ground conductivity) the direction finder approaches one terrestrial normally-polarized wave. Error-free direction finding of short waves it is possible to carry out, applying any antenna systems (framework, the spaced antennas, the diverse framework). It is most expedient in these distances to apply frame antenna system.

From the ionosphere the wave reflected can appear either at certain distance, after will virtually disappear terrestrial wave or

when terrestrial wave still it does not decrease so, so that it would be possible it not to consider. In the first case there is a dead spot, in the second case will appear the section where simultaneously there are terrestrial and reflected waves.

The limits of dead spot for any wave depend on time of days and year, on the power of transmitter and parameters of soil. The greater the ground conductivity, the further is propagated the terrestrial wave and is less dead spot.

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In dead spot both reception/procedure and the direction finding are unreliable. There can exist only diffuse reflections with direction finding of which are obtained the inaccurate bearings of the changing value, any in no way is obtained bearing. In dead spot, all antenna systems are equally unreliable, since the direction of the arrival of the scattered ray/beams has usually no relation to real direction in radio station.

At those distances where there is terrestrial and sky wave and where sky wave composes only several percentages from terrestrial wave, and consequently, there are no polarizational errors, it is possible to utilize any system, including simplest - frame. At large

distances, where sky wave grow/rises, it is necessary to apply the antenna system, free from polarizational errors.

When selecting by the antenna of system, free from polarization errors, one should proceed from the fact that at abrupt/steep angles of incidence ($\beta > 60$-$65°$) the quality of the work of system with the spaced antennas considerably deteriorates. This is explained to the facts that relationship/ratio between the vertical and horizontal components of electric field, proportional $\cos^2 \beta$, decreases in comparison with propagation along the horizon/level:

— при(1) $\beta = 60°$ $\cos^2 \beta = 0{,}25$ (ухудшение(2) в 4 раза),
— при $\beta = 80°$ $\cos^2 \beta = 0{,}03$ (ухудшение более чем (3) в 30 раз).

Key: (1). with. (2). deterioration 4 times. (3). deterioration is more than 30 times.

This deficiency/lack, as we see that does not have any the system with the diverse framework, since the reception/procedure vertical component this system does not depend on β, but the reception/procedure of the horizontal component of electric field is proportional to $\cos \beta$. As a result of this, the relationship/ratio between voltage/stresses from horizontal component and from vertical component electric field decreases with an increase in the angle of

incidence. Only where the angle of incidence lie/rests within limits approximately 0-45°, one should apply system with the spaced antennas. This corresponds to distances from transmitter, to high approximately 400 km. At the distances where this angle ranges from 45 to 60-65° (ranges 400-600 km), a system with the diverse framework has small advantages over the system with the spaced antennas and finally at angles of incidence 60-65° to 90° (range to 400 km) must be applied system with the diverse framework.

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One should emphasize that system with the diverse framework it is expedient to apply only in those distances where it has explicit advantages on accuracy in comparison with system with the spaced antennas, since system with the diverse framework is less sensitive and bulkier.

From Fig. 6.26, which represents the dependence of distance from angle of incidence for reflections from layers E, F_1 and F_2, it is possible to establish/install for the specific conditions of propagation those ranges at which it is expedient to use one or the other antenna system in the radio direction finders of short waves. The indicated in figure ranges are some average.

In night hours the relationship/ratio between the vertical and horizontal components of electric field deteriorates in comparison with daytime hours. Therefore, as a rule, at night polarizational errors appear more powerful than in the daytime, and also increase lateral divergences.

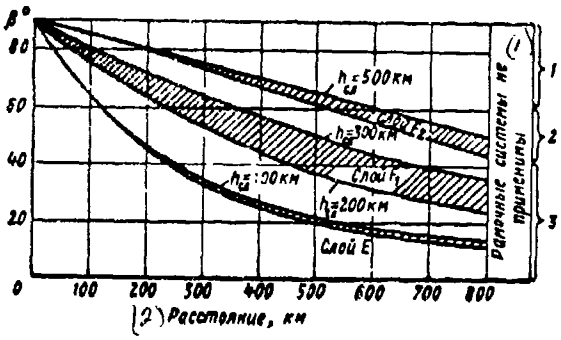

Fig. 6.26. Dependence of angle of incidence on distance: 1 - diverse framework are much better than spaced antennas; 2 - diverse framework are somewhat better than spaced antennas; 3 - diverse framework and separated antennas have approximately identical accuracy.

Key: (1). Frame systems are not used. (2). Distance, km.

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At the very large distances, close to antipodes, the direction finding becomes impossible. On the system, free from polarizational errors which only and it is possible to use here are observed the changing in time readings. This is explained to the facts that from the numerous emission/radiations, passed approximately equidistances

of direction finder, most intense prove to be the electromagnetic vibrations, passed, although somewhat longer path, corresponds to this direction, which is changed in the course of twenty-four hours.

Since the operating errors of radio direction finder to a considerable degree depend on the susceptibility of radio direction finder to polarizational errors, the problem of a decrease in the standard polarizational error is extremely urgent. Simultaneously with this stands the problem of a decrease in the errors due to lateral divergences and interference. For an increase in the accuracy of direction finding where this is attained, one should apply antenna system with the large separation of antennas or utilize several diverse radio direction finders (2-3) for the direction finding of just one transmitter. Direction finders must be spread up to the distance, greater than 200-300 m. Furthermore that in this way are averaged errors with direction finding, it is possible according to the character of readings on separate equipment/devices and by a difference in the readings to judge the conditions of direction finding and the reliability of bearings. It is very useful to available in simultaneously working radio direction finders different external systems - with the spaced antennas and the diverse framework. Theoretically mean error must with this direction-finding method vary in proportion to $\frac{1}{\sqrt{n}}$, where n - a number is simultaneous of working identical radio direction finders.

Direction finding on ultra short waves (30-300 MHz).

At these frequencies the direction finding is realized on terrestrial wave. The direction finding of utilized dol. of the radio communication of the scattered from the ionosphere and the troposphere waves is very difficult. The range of transmitter depends on the height/altitude of the antenna location of transmitter and receiver.

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At the small height/altitudes of the location of the transmitting antenna, the strength of field from the vertical wire antenna is more than from horizontal, since the horizontal component of electric field rapidly attenuates at the earth/ground; at the high altitude of the location of the transmitting antenna vice versa. Thus, at these frequencies are possible the waves of any polarization.

Within the limits of the city where there are many structures, and also in broken ground to receiving antenna, comes the resulting

field from the transmitting antenna and from the reflecting object/subjects. Therefore the direction of arrival and the polarization of wave can be connected with direction and polarization of the transmitting antenna. The inequalities of soil cause an increase in the absorption and a decrease in the transmission distance. At distances, large of direct/straight visibility, is developed the effect of the troposphere.

On ultra short waves are predominantly applied the radio direction finders with rotatable loop or rotary pair of antennas either greater base radio direction finders with Circular or line-source antenna by system from the spaced antennas. Sometimes in the rotary pair of antennas, is provided for the possibility of a change in the orientation of antennas for the reception/procedure of the field of vertical and horizontal polarization.

In by an antenna to system with motionless antennas are utilized the following principles of the reading of the bearing: two-channel, phase with the cyclic measurement of phase in high frequency, with scanning of acute/sharp directional characteristic, etc. When selecting by the antenna of system it is necessary to consider that with an increase in the separation of antennas decreases the influence of environment, which very is substantial for the radio direction finders, working on ultra short waves.

Chapter 7.

CALCULATION OF ANTENNA SYSTEMS OF RADIO DIRECTION FINDERS.

7.1. Preliminary considerations.

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The fundamental characteristics of radio direction finder are its accuracy and sensitivity.

Accuracy is determined by the average quadratic (or average arithmetic) value of the random angular error of radio direction finder. The components of the random angular operating error of radio direction finder are examined in §§2.4, 4.1-4.12, 5.3, 6.1-6.5 and 11.1. In the design of radio direction finder, after are selected the principle of direction finding, the method of the reading of bearing and the used antenna system, is designed or is rate/estimated each

component error and is determined the total random error of radio direction finder.

The sensitivity of radio direction finder is measured by the strength of field, which is required for providing the assigned subjective error of the reading of bearing. last/latter it causes the required for the reading of bearing sense of the voltage/stresses of signal and noise on the input of receiving indicator. Thus, sensitivity - this is that minimum strength of field with which is obtained the necessary sense of the voltage/stresses of signal and noise on the input of receiving indicator.

As it is shown into §2.5, the sensitivity of radio direction finder is improved with a decrease in the factor of noise N of the input part of reception indicator, a decrease in its passband and with growth in the product $D\eta$, where D is a directive gain of antenna, η— the efficiency of antenna feeder circuit.

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After replacing D by its expression (see §2.5) and after considering that $\eta=\eta_{n}\eta_{\phi}$, where η_{n}— radiation efficiency, η_{ϕ}— the efficiency of the feeder, which connects antenna with the input of receiving indicator, let us explain that for an improvement in the

sensitivity it is necessary to increase the parameter by the antenna of the system

$$B_a = \frac{H_e^2}{R_a} \eta_\Phi,$$

where H_e and R_a — effective height and effective resistance by the antenna of system.

With the auditory method of the reading of bearing on the minimum for an improvement in the sensitivity, it is necessary to also attain an increase in the slope/transconductance of the differential characteristic of directivity at zero. An increase in this slope/transconductance is required also with other methods of reading (sum-and-difference, with modulation on input, etc.).

Factor of noise N depends on the transmission gain from antenna on the input of reception indicator. Taking into account the small difference of the optimum factor of noise and noise factor with the maximum transmission gain of input circuits of receiving indicator, initial calculation of input circuit we produce, on the basis of the condition of obtaining the maximum of the transmission of voltage/stress (§§7.3-7.10). Subsequently noise factor, it can be checked and the corresponding values are corrected (§ 7.2).

During the calculation of input device, it is necessary to also

consider the conditions of the tuning of input circuits of receiving indicator. In the majority of contemporary receivers, the tuning of input circuits is united with the tuning of all remaining circuits of receiver. Therefore the detuning of the first duct of receiver under the effect of antenna coupling system must not be great, which limits the permissible communication/connection of input circuit with antenna system.

For an increase in the parameter by the antenna of system B_a it is necessary to obtain a maximally possible value for $\frac{H_e^2}{R_a}$ and for η_c Maximum $\frac{H_e^2}{R_a}$ they reach by the appropriate selection of size/dimensions and constructions of separate antennas and by entire antenna of system (by the appropriate selection of the turn number of the framework, by the correct selection of diameter and height/altitudes of separate vibrators and construction made of groundings, by the selection of separation and numbers of antennas of system, etc.).

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The efficiency of feeder $\eta_ф$ will be the greater, the nearer to one KBV of feeder. Equality KBV = 1 corresponds to the condition of the matched load, when load impedance is equal to the wave impedance of feeder. It should be noted that when KBV = 1 considerably is

facilitated the design of the input part of the receiving indicator, since in this case it is easy to compensate for reaction by the antenna of system to inclined input circuit and to obtain the optimum value of noise factor on all working frequency band. In §4.12 shown, that when KBV = 1, the possible dissimilarity of the parameters of high-frequency cables is led to minimum instrument errors. Is explained this to the facts that when KBV = 1 there are no resonances in antenna feeder system and the dissimilarity of cell/elements and the possible asymmetry of antennas is not created large errors.

Thus, are obvious the positive qualities of the agreement of the loads of the high-frequency cables, which connect antennas to receiving indicator, with the wave impedance of cable. This is especially expedient, when antenna system is arrange/located at a great distance, equal to several wavelengths.

Therefore during the design of antenna feeder system, it is necessary to examine the possibility of the selection of such construction of antennas during which is provided in all working frequency band antenna matching and feeder. For this, as is known, are applied large-diameter vertical wire antennas (with small we will use vertical wire antennas with the thrown in cell/elements L and P (§3.1) and of the diagram of the compensation for the reactance of vibrators with the aid of the concentrated cell/elements or the

cuts of long line and, etc.

When it is not represented possible to achieve agreement by the selection of construction or antenna circuits, sometimes between feeder and load, connect the special adapters, serving for the transformation of the impedance of antenna into the effective resistance, equal to the wave impedance of feeder. Finally, if the switching on of the cell/elements of agreement is impossible or it is inexpedient, is applied the mismatched antenna feeder system.

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In this case it is desirable not to have resonances in system in working frequency band and to obtain KBV of feeders as close as possible to unity. Sometimes in antenna feeder system it is necessary to switch on supplementary effective resistance in order to decrease the manifestation of resonances, if cannot be avoided them.

From that which was presented it follows that are possible three versions of the calculation of antenna feeder system and the input device of the receiver of the radio direction finder:

1. Direct connection of feeders to the mismatched with feeders loads (antennas) and obtaining maximum attainable in the working

frequency band of the transmission factor and sensitivity. In this case, it is necessary to consider the effect of the reaction of the resistor/resistance of antenna feeder system for the tuning of input circuit; must be accepted measures to the limitation of instrument errors due to the possible dissimilarity of the cell/elements of system and asymmetry of separate antennas. Calculations are given into §§7.3-7.8.

2. Application/use of cell/elements of agreement between antenna and feeder. Matching device is necessary to select, taking into account the law of change with the frequency of load impedance. To usually previously solve the question concerning the advisability of applying the adapters of the agreement of loads is not represented possible. This question is solved as a result of the corresponding calculations.

3. Sometimes is possible impedance matching of antenna with feeder without application/use of supplementary cell/elements. This corresponds to the use of wide-range vibrators (§3.1). The advisability of making one or the other decision must be determined as a result of design.

7.2. Coefficient of the noise of input circuit.

The majority of diagrams examined below of input circuits can be given to the common equivalent diagram, presented in Fig. 7.1.

Let us designate:

X_a, R_a — the jet/reactive and effective resistance of antenna (converted to the input of receiver);

L_1, R_1 are inductance and the effective resistance of coupling coil;

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C_2, L_2, R_2 are capacitance/capacity, the inductance and the effective resistance of resonant circuit;

R_n — noise lamp resistance;

K - the coupling coefficient of the coils of duct and communication/connection.

Let us find the resistor/resistances, introduced by primary circuit in secondary:

$$\Delta X = -\frac{\omega^2 M^2}{z_1^2}(X_a + X_1) = -n^2 (X_a + X_1), \qquad (7.1)$$

$$\Delta R = \frac{\omega^2 M^2}{z_1^2}(R_a + R_1) = n^2 (R_a + R_1), \qquad (7.2)$$

where

$$z_1 = \sqrt{(X_a + X_1)^2 + (R_a + R_1)^2};\ X_1 = \omega L_1;\ n = \frac{K\omega \sqrt{L_1 L_2}}{z_1}.$$

For reduction let us introduce the designations

$$\omega L_2 - \frac{1}{\omega C_2} - n^2 (X_a + X_1) = X',$$
$$4kTB = P_0.$$

If one assumes that the sole noise source is antenna resistance, then the square of the effective value of noise voltage in secondary duct will be

$$E_{шa}^2 = P_0 n^2 R_a,$$

and the square of noise voltage on the terminal/grippers of the duct

$$U_{шa}^2 = \frac{E_{шa}^2}{(X')^2 + (R_2 + \Delta R)^2} \frac{1}{\omega^2 C_2^2} = \frac{P_0 n^2 R_a}{(X')^2 + (R_2 + \Delta R)^2} \frac{1}{\omega^2 C_2^2}.$$

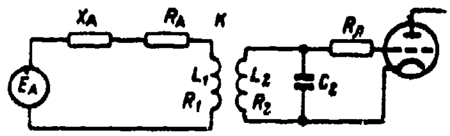

Fig. 7.1. Equivalent diagram of input circuit.

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If we consider the action of all noise sources, then the square of complete noise voltage on the grid of the first tube will be expressed by

$$U_{ш}^2=\frac{P_0\,(n^2R_A+n^2R_1+R_2)}{[(X')^2+(R_2+\Delta R)^2]\,\omega^2C_2^2}+P_0R_{ш}. \qquad (7.3)$$

Hence we determine noise factor:

$$N=\frac{U_{ш}^2}{U_{шA}^2}=1+\frac{R_1}{R_A}+\frac{R_2}{n^2R_A}+R_{ш}\,\frac{[(X')^2+(R_2+\Delta R)^2]\,\omega^2C_2^2}{n^2R_A}. \qquad (7.4)$$

Let us examine the case, when noise lamp resistance is very small, so that last/latter term in expression for a noise factor can be disregarded.

The factor of the noise of Budde then is equal to

$$N = 1 + \frac{R_1}{R_a} + \frac{R_2}{n^2 R_a}. \qquad (7.5)$$

From this expression it is evident that for a decrease in the noise factor it is necessary, as far as possible, to increase n. This can be reached by an increase in the coupling coefficient K and by the appropriate selection of value L_1. Hence it follows that K must be selected by as large, as this is possible structurally. As concerns the inductance of coupling coil, then with its increase change both numerator and the denominator of expression for n. Furthermore, during practical fulfillment simultaneously with loading in the coupling coil increases its effective resistance. Usually it is possible to consider that the coil Q of communication/connection has constant value Q_1:

$$R_1 = \frac{\omega L_1}{Q_1}.$$

Taking into account these facts, we will find the most advantageous value L_1. For this purpose, let us write the expanded/scarned expression for n^2:

$$n^2 = \frac{\omega^2 K^2 L_1 L_2}{(X_a + \omega L_1)^2 + \left(R_a + \frac{\omega L_1}{Q_1}\right)^2}$$

let us substitute it into formula (7.5):

$$N = 1 + \frac{\omega L_1}{Q_1 R_a} + \frac{R_2}{R_a} \frac{(X_a + \omega L_1)^2 + \left(R_a + \frac{\omega L_1}{Q_1}\right)^2}{\omega^2 K^2 L_1 L_2}.$$

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Sweeping this expression, disregarding small term $(\omega L_1)/Q_1)^2$ and banking terms with identical degrees ωL_1, we will obtain

$$N = 1 + \frac{R_2}{K^2 \omega L_2}\left\{2\left(\frac{X_a}{R_a} + \frac{1}{Q_1}\right) + \right.$$
$$\left. + \omega L_1\left(1 + \frac{K^2 \omega L_2}{Q_1 R_2}\right)\frac{1}{R_a} + \frac{1}{\omega L_1}\frac{X_a^2 + R_a^2}{R_a}\right\}. \quad (7.6)$$

The minimum of noise factor is obtained, when

$$(\omega L_1)_{\text{опт}} = \sqrt{\frac{X_a^2 + R_a^2}{1 + \frac{K^2 \omega L_2}{Q_1 R_2}}} = \frac{z_a \sqrt{Q_1}}{\sqrt{K^2 Q_2 + Q_1}}, \quad (7.7)$$

where

$$z_a = \sqrt{R_a^2 + X_a^2};\ Q_2 = \frac{\omega L_2}{R_2}.$$

The substitution of this value $\omega L_1 = \omega L_{1\text{опт}}$ in (7.6) gives the minimum value of factor of the noise

$$N_{\text{мин}} = 1 + \frac{2}{K^2 Q_2}\left\{\left(\frac{X_a}{R_a} + \frac{1}{Q_1}\right) + \frac{z_a}{R_a}\sqrt{\frac{K^2 Q_2 + Q_1}{Q_1}}\right\}. \quad (7.8)$$

From last/latter formula it is evident that the coupling coefficient must be undertaken maximum (this noted earlier than). Furthermore, it is necessary to ensure the maximum energy factor Q_2

of resonant circuit and quality Q_1 of coupling coil, although the latter affects the value of noise factor considerably weaker than the energy factor of resonant circuit and coupling coefficient.

If we by a designate the relation

$$a = \frac{L_{1\,\text{опт}}}{L},$$

then noise factor can be represented in the form

$$N = 1 + \frac{1}{K^2 Q_2}\left\{2\left(\frac{X_a}{R_a} + \frac{1}{Q_1}\right) + \frac{z_a}{R_a}\sqrt{\frac{K^2 Q_2 + Q_1}{Q_1}}\left(a + \frac{1}{a}\right)\right\} = \\ = N_{\text{мин}} + \frac{z_a}{R_a}\sqrt{\frac{K^2 Q_2 + Q_1}{Q_1}}\,\frac{(a-1)^2}{a}. \qquad (7.9)$$

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Last/latter formula makes it possible to rate/estimate the effect of a difference in the inductance of coupling coil from optimum value on the value of noise factor.

Let us examine the application/use of the obtained formulas to different type antennas, by the differing character of reactance.

1. Reactance of antenna we have inductive character (framework).

$$X_a = \omega L_a,$$

$$(\omega L_1)_{\text{опт}} = \frac{\sqrt{[(\omega L_a)^2 + R_a^2]\,Q_1}}{\sqrt{K^2 Q_2 + Q_1}}. \qquad (7.10)$$

Formula (7.10) shows that the inductance of coupling coil must be less than the inductance of the framework.

According to formula (7.8), disregarding value $1/Q_1$ in comparison with $\frac{X_a}{R_a}=Q_a$, we will obtain for a minimum factor of the noise

$$N_{\text{мин}}=1+\frac{2Q_a}{K^2Q_2}\left\{1+\sqrt{\frac{K^2Q_2}{Q_1}+1}\right\}. \qquad (7.11)$$

On Fig. 7.2, is represented the graph/diagram of the dependence $N_{\text{мин}}$ on $\frac{K^2Q_2}{Q_a}$ for the case when $Q_1=Q_a$.

2. Reactance of antenna is equal to zero. The optimum inductance of coupling coil and minimum noise factor will be

$$(\omega L_1)_{\text{опт}}=\frac{R_a}{\sqrt{\frac{K^2Q_2}{Q_1}+1}}, \qquad (7.12)$$

$$N_{\text{мин}}=1+\frac{2}{K^2Q_2}\sqrt{\frac{K^2Q_2}{Q_1}+1}. \qquad (7.13)$$

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3. Reactance of antenna has capacitive character. Disregarding the effective resistance of antenna, we will obtain optimum inductance of coupling coil from

$$(\omega L_1)_{опт} = \frac{1}{\omega C_a} \sqrt{\frac{Q_1}{K^2 Q_2 + Q_1}}. \qquad (7.14)$$

This expression determines the optimum tuning of the antenna

$$(\omega^2 L_1 C_a)_{опт} = \sqrt{\frac{Q_1}{K^2 Q_2 + Q_1}}. \qquad (7.15)$$

Minimum noise factor will be

$$N_{мин} = 1 + \frac{2Q_a}{K^2 Q_2}\left[1 + \sqrt{\frac{K^2 Q_2 + Q_1}{Q_1}}\right]. \qquad (7.16)$$

From expression (7.15) it is evident that the resonance frequency the antenna of circuit in this case must be higher than operating frequency.

The obtained formulas make it possible to calculate the cell/elements of input circuit of receiving indicator, if is known antenna resistance at the input of receiving indicator and it is possible to disregard the noise resistor/resistance of input time. Since antenna resistance does not remain constant over a wide range of frequencies, the formulas of present paragraph can be applied only for the calculation of the cell/elements of input with narrow

frequency band when antenna resistance can be considered constant, and also for a comparison at the various frequencies of the operating range of minimum coefficient of noise with coefficient of noise, obtained as a result of the calculations, given in the subsequent paragraphs. These calculations are carried out on the basis of the requirement for obtaining the maximum transmission factor of input circuit.

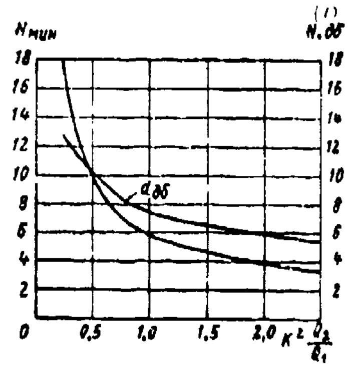

Fig. 7.2. Factor of the noise of frame direction finder.

Key: (1). dB.

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7.3. Calculation of the effectiveness of spiral loop.

Simplest diagram of the framework is shown in Fig. 7.3.

The framework, which has the operating inductance L, and effective resistance R_0, is tuned with the aid of condenser/capacitor C. If we, designate emf, induced within the framework, by E, then

with resonance, when

$$\omega L_0 = \frac{1}{\omega C},$$

current $I_{рез}$ in the duct of the framework is equal to

$$I_{рез} = \frac{E}{R_0}.$$

Voltage on condenser/capacitor, supplied to the gril of the first tube, will be

$$U_{c\,рез} = \frac{I_{рез}}{\omega C} = \frac{E}{\omega C R_0}. \qquad (7.17)$$

Ratio $1/\omega CR_0 = \omega L_0/R_0 = Q$ is the quality of the framework. Thus, for the transmission gain the simple spiral loop we will obtain

$$k_п = \frac{U_{c\,рез}}{E} = Q. \qquad (7.18)$$

The effectiveness of spiral loop, i.e., the ratio of grid voltage of the first tube $U_{c\,рез}$ to the strength of field E will be

$$p_п = \frac{U_{c\,рез}}{E} = h_д Q = \frac{2\pi S N_p}{\lambda} Q \ м, \qquad (7.19)$$

where S is an area of the framework, m²;

λ - wavelength, m;

N_p -- turn number.

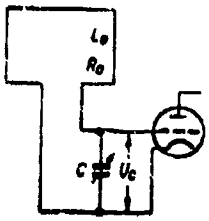

Fig. 7.3. Circuit diagram of spiral loop.

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In the balanced network of the switching on of the framework (Fig. 7.4) on the grid of tube falls only the half of the voltage, which is received on condenser/capacitor. Consequently, effectiveness will be

$$p_n = \frac{h_r Q}{2}.$$

It should not be supposed that the effectiveness of the balanced network is two times lower than asymmetric one. The fact is that R_0 in the given above formulas one should understand as effective resistance, equivalent to all losses in framework and the connected with it circuits, including resistor/resistance, equivalent to losses in the circuit of the grid of tube. With the balanced network of the inclusion this component/term of resistor/resistance decreases 4

times, which leads to an increase in the factor of resonance and, therefore, effectiveness.

If necessary to overlap wide frequency band is possible the application/use of the subdivided framework. Transition from one partial range on another is realized by switching on of the larger or smaller turn number of the framework. The inoperative (idle) turns of the framework remain in this case extended or are closed short. A deficiency/lack in this method lies in the fact that the presence of dead turns can cause the asymmetry of the framework, therefore, an increase in the antenna effect. Furthermore, the presence of dead turns leads to an increase in the attenuation, i.e., to a fall in quality and effectiveness of the framework.

Another method of the overlap of wide wave band consists of the application/use of diagrams of shortening (Fig. 7.4a) and of the elongation of the wave of the framework (Fig. 7.4b). In the diagram of shortening in parallel to the framework is included the self-inductor L_{H}.

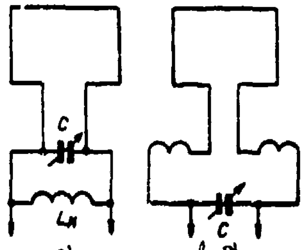

Fig. 7.4. Diagrams of the wave band of the framework: a) shortening; b) elongation.

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The resulting inductance therefore decreases and the framework is tuned to shorter wave. Equivalent diagram for the calculation of effectiveness in the case of the shortening of the wave of the framework is shown in Fig. 7.5. Let us designate impedance between points a and b (to the right of them) by Z_{ab}:

$$Z_{ab}=\frac{R_н+j\omega L_н}{1-\omega^2 L_н C+j\omega R_н C}.$$

Voltage across capacitor U_c will be equal

$$\dot{U}_c=E\frac{Z_{ab}}{Z_{ab}+R_e+j\omega L_e}=$$

$$=\frac{E\,(R_н+j\omega L_н)}{R_н(1-\omega^2 L_e C)+R_e\,(1-\omega^2 L_н C)+j\omega(L_н+L_e+R_н R_e C-\omega^2 L_н L_e C)}.$$

Voltage across capacitor maximum, when the imaginary part of the denominator is turned into zero. Disregarding low value $R_{\kappa}R_0C$ in comparison with the others, resonance condition we will obtain in the form

$$\omega=\sqrt{\frac{L_{\kappa}+L_{0}}{L_{\kappa}L_{0}C}}=\sqrt{\frac{1}{L'C}}, \qquad (7.20)$$

where $L'=\frac{L_{\kappa}L_{0}}{L_{\kappa}+L_{0}}$ — the inductance, which consists of the parallel-connected inductance of the framework and shortening coil.

After supplying this condition in expression for U_c and disregarding R_{κ} in comparison with ωL_{κ} in numerator, let us find voltage across capacitor at the resonance:

$$U_{c\,res}=E\frac{\omega L_{0}}{R_{0}+\frac{L_{0}^{2}}{L_{\kappa}^{2}}R_{\kappa}}=EQ'. \qquad (7.21)$$

In formula (7.21) Q' is an equivalent energy factor of duct. It is

obvious that Q' are less than Q.

The transmission gain of the framework is equal to

$$k_{n}=Q'=Q\frac{1}{1+\frac{L_0^2 R_{K}}{L_{K}^2 R_0}}=\frac{Q}{1+\frac{L_0 Q}{L_{K} Q_{K}}}, \qquad (7.22)$$

where Q_K — the quality of the shortening coil.

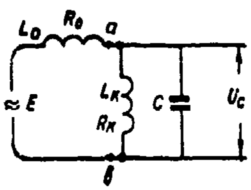

Fig. 7.5. Equivalent diagram of the framework with shortening.

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Effectiveness p_0 is equal to

$$p_0=h_c Q'.$$

In the case of the elongation of the wave of the framework (Fig. 7.4b) can be without change used formula (7.18) of the transmission gain. One should only consider that hearth R_0 here is implied the total resistance of the framework and coil. Although the transmission gain is obtained much the same, as for circuit Fig. 7.3, in this case

effectiveness is obtained lower, since the turn number of the framework is taken smaller than this would be possibly during application/use on the same wave of the diagram of adjusted framework without elongation.

Let us pause at the question concerning the selection of the fundamental parameters of the framework - its diameter (or side) and turn numbers. From the viewpoint of obtaining the greatest effectiveness to favorably make the linear dimensions of the framework largest possible. Therefore them one should select by as large, as this allow practical considerations - convenience in the arrangement/permutation and rotation. With the assigned linear dimensions of the framework for a work on one fixed/recorded wave, it is possible to fit most advantageous turn number. The existance of optimum for a turn number is determined by the facts that the effective height of the framework with an increase in the turn number grow/rises, and the transmission gain beginning with certain turn number falls as a result of a sharp incidence/drop in quality [7.1].

Virtually always problem of work with one framework in certain frequency band. In that case the turn number of the framework is determined by its inductance which is assigned by capacitance value of alternating/variable condenser/capacitor and by the selected diagram of input circuit. It should be noted that the framework and

especially the wires, which connect the framework with receiver, possess a comparatively great capacity. Because of this maximum capacitance of alternating/variable capacitor is required greater than for the overlap of the same wave band in usual ducts.

If necessary to overlap wide frequency band when it is required to relate the framework of receiver. expedient to use the diagram of inductive coupling with the unadjusted framework.

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On the calculation of this diagram, is reduced also the calculation of frame system from goniometer. This calculations are given into §7.4.

7.4. Calculation of the effectiveness of the framework with inductive coupling.

Unadjusted framework with inductive coupling.

In this case (Fig. 7.6), disregarding the effect of the

self-capacitance of the framework, its impedance we will obtain in the form

$$Z_1 = R_0 + R_1 + j\omega(L_0 + L_1).$$

Circuital current of the framework is equal to

$$\dot{I}_1 = \frac{\dot{E}_1}{R_0 + R_1 + j\omega(L_0 + L_1)}.$$

The introduced into secondary duct resistor/resistance is equal

$$\Delta Z_2 = \frac{\omega^2 M^2}{Z_1} = \frac{\omega^2 M^2}{(R_0 + R_1)^2 + \omega^2 (L_0 + L_1)^2}[(R_0 + R_1) - \\ - j\omega(L_0 + L_1)],$$

where M is the mutual inductance of the lucts of the framework and input (secondary):

$$M = K\sqrt{(L_0 + L_1) L_2}.$$

Total impedance of secondary circuit is equal

$$Z'_2 = Z_2 + \Delta Z_2 = R_2 + \frac{\omega^2 M^2 (R_0 + R_1)}{(R_0 + R_1)^2 + \omega^2 (L_0 + L_1)^2} + \\ + j\left[\omega L_2 - \frac{1}{\omega C_2} - \frac{\omega^3 (L_0 + L_1) M^2}{(R_0 + R_1)^2 + \omega^2 (L_0 + L_1)^2}\right].$$

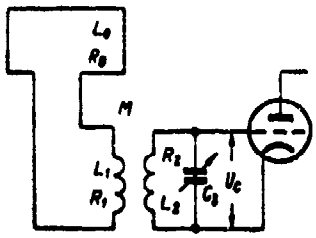

Fig. 7.6. Circuits of the switching on of the framework with inductive coupling.

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In the presence of resonance the imaginary part of this expression is turned into zero:

$$\omega L_2 - \frac{1}{\omega C_2} - \frac{\omega^2 (L_0 + L_1) M^2}{(R_0 + R_1)^2 + \omega^2 (L_0 + L_1)^2} = 0.$$

Since $R_0 + R_1$ usually is considerably less than $\omega(L_0 + L_1)$, in the denominator of last/latter term it is possible to disregard first term. Then we will obtain the approximate resonance condition

$$\omega\left(L_2 - \frac{M^2}{L_0 + L_1}\right) - \frac{1}{\omega C_2} = 0.$$

We see that the effect of the core circuit, equivalent to a decrease in the inductance of secondary, by value

$$\Delta L_2 = \frac{M^2}{L_0 + L_1} = K^2 L_2 = K_L^2 \frac{L_1 L_{\text{св}}}{L_0 + L_1},$$

where $L_{cв}$ — the part of coil L_2, connected with L_1,

$$M = K_L \sqrt{L_{cв} L_1}.$$

Since this change does not depend on frequency [1], it easily can be compensated for corresponding increase in its own inductance of the secondary duct L_2.

FOOTNOTE [1]. This is correct only neglecting of the capacitance/capacity of the framework, i.e., when $\lambda_0 < \lambda_{мин}$. ENDFOOTNOTE.

The resistor/resistance of secondary circuit in the presence of resonance is equal

$$Z'_{2\,рез} = R_2 + \frac{M^2}{(L_0 + L_1)^2}(R_0 + R_1) = R_2 + \\ + K_L^2 \frac{L_1 L_{св}}{(L_0 + L_1)^2}(R_0 + R_1).$$

During the derivation of this formula, we also disregard member $(R_0 + R_1)^2$ in comparison from $\omega^2 (L_0 + L_1)^2$.

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Let us determine now voltage across capacitor in the presence of the resonance:

$$\dot U_{c\,\text{рез}} = -j\frac{\dot E_2}{Z'_{2\,\text{рез}}}\frac{1}{\omega C_2} = \frac{\omega^2 M}{Z'_{2\,\text{рез}}}\left(L_2 - \frac{M^2}{L_0+L_1}\right)\dot I_1 = \\ = \frac{\omega^2 M\left(L_2 - \frac{M^2}{L_0+L_1}\right)}{Z'_{2\,\text{рез}}}\frac{E_1}{(R_0+R_1)+j\omega(L_0+L_1)}.$$

Again, disregarding $(R_0 + R_1)^2$ in comparison from $\omega^2(L_0 + L_1)^2$, we find the amplitude of voltage across capacitor:

$$U_{c\,\text{рез}} = \frac{\omega M\left(L_2 - \frac{M^2}{L_0+L_1}\right)}{\left[R_2 + \frac{M^2}{(L_0+L_1)^2}(R_0+R_1)\right](L_0+L_1)}E_1. \quad (7.23)$$

We hence find the transmission gain

$$k_0 = \frac{\omega M\left(L_2 - \frac{M^2}{L_0+L_1}\right)}{\left[R_2 + \frac{M^2}{(L_0+L_1)^2}(R_0+R_1)\right](L_0+L_1)}. \quad (7.24)$$

We convert last/latter formula. Introducing designation $\alpha = L_1/L_0$ and taking into account the equality

$$K_L = \frac{M}{\sqrt{L_1 L_{\text{св}}}},\ \delta_1 = \frac{R_0+R_1}{\omega(L_0+L_1)},\ \delta_2 = \frac{R_2}{\omega L_2},$$

we will obtain

$$k_n = \frac{\sqrt{\alpha}}{1+\alpha}\,\frac{K_L\left(1 - K_L^2\frac{\alpha}{1+\alpha}\frac{L_{\text{св}}}{L_2}\right)}{\delta_2 + K_L^2\frac{\alpha}{1+\alpha}\frac{L_{\text{св}}}{L_2}\delta_1}\sqrt{\frac{L_{\text{св}}}{L_1}}, \qquad (7.25)$$

where δ_1, δ_2 - circuit damping of the framework and input.

Differentiating this expression on α and equalizing zero for the target/purpose of the determination of maximum k_n, we obtain complete cubic equation relatively α. Taking into account that near maximum value k_n varies little, for simplification in the calculation admissibly to find optimum value α, after placing in the last/latter formula

$$1 - K_L^2\frac{\alpha}{1+\alpha}\frac{L_{\text{св}}}{L_2} \approx 1.$$

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It is really/actually

$$K_L^2\frac{\alpha}{1+\alpha}\frac{L_{\text{св}}}{L_2} \ll 1.$$

Under these assumptions

$$\alpha_{\text{опт}} = \frac{\delta_2}{\delta_2 + K_L^2\frac{L_{\text{св}}}{L_2}\delta_1} \qquad (7.26)$$

and

$$k_{\text{a макс}} = \frac{K_L}{2\delta_2}\sqrt{\frac{\delta_2}{\delta_2 + K_L^2\delta_1\frac{L_{\text{сп}}}{L_2}}} \times$$

$$\times \frac{(2-K_L^2)\,\delta_2 + K_L^2\delta_1\frac{L_{\text{сп}}}{L_2}}{2\delta_2 + K_L^2\delta_1\frac{L_{\text{сп}}}{L_2}}\sqrt{\frac{L_{\text{сп}}}{L_0}} \qquad (7.27)$$

On Fig. 7.7, is given the graph/diagram of the dependence $k_{\text{a макс}}$ on K_L at the different values of the ratio δ_1/δ_2 and when $L_{\text{сп}}=L_2$. From curve/graph it is evident that $k_{\text{a макс}}$ grow/rises almost proportionally K_L to the values K_L, approximately equal to 0.5-0.6. With a further increase K_L an increase in the effectiveness occurs slowly.

Let us examine considerations by choice of the parameters of the framework for the case of the unadjusted framework. From formula (7.27) it is evident that $k_{\text{a макс}}$ the unadjusted framework inversely proportional to square root of its inductance.

It is concealed by shape, the effectiveness of system is proportional to value

$$\frac{2\pi SN_p}{\lambda}\cdot\frac{1}{\sqrt{L_0}}.$$

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Consequently, the area of the framework it is advantageous to make largest possible.

When selecting turn number N_p one should consider that value L_0 also depends on N_p. This dependence takes the following form:

$$L_0 = A + BN_p + CN_p^2,$$

where A, B and C - the coefficients, determined by size/dimensions and the form of the framework.

Value $\frac{N_p}{\sqrt{L_0}} = \frac{N_p}{\sqrt{A + BN_p + CN_p^2}}$ at small turn number grow/rises almost proportionally N_p, and then with large N_p it approaches limit $\lim\left(\frac{N_p}{\sqrt{L_0}}\right) = \frac{1}{\sqrt{C}}$.

A considerable increase N_p can lead to the fact that its own wave of system will render/show within working wave band, as a result of which appears nonuniform effectiveness in range and the whole series of complications during the use of the framework in radio direction finder. Furthermore, in this case are not used the derived above formulas (since we are disregarded C_0).

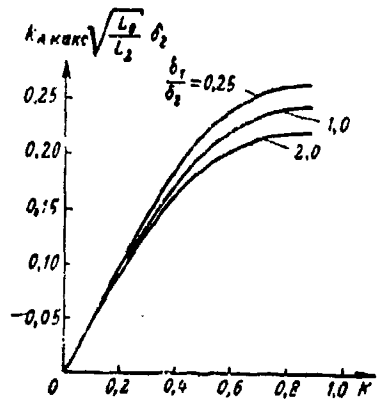

Fig. 7.7. Dependence $k_{A\,макс}$ on K_L for a spiral loop with inductive communication/connection.

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Therefore they are usually satisfied by achievement 90-95o/o of limiting value $\frac{N_r}{\sqrt{L_0}}$, limiting at the same time its own wave of the framework (taking into account the capacitance/capacity of lead wires) by value (0.5-0.[illegible]) $\lambda_{мин}$.

The unadjusted framework possesses somewhat smaller effectiveness than inclined. The loss in the application/use of inductive coupling the greater, the higher the quality of the framework. On the other hand, the application/use of inductive

coupling provides the large uniformity of effectiveness during band selection of waves and therefore it gives favorable results with the very wide wave band, which requires the large number of switchings of partial ranges.

Goniometric system with the locked framework.

Effectiveness goniometric of system of two mutually perpendicular framework is equal to the effectiveness of rotatable loop with the inductive coupling which has the same diagram of input as in goniometric system with the maximum communication/connection of the coils of goniometer. On the basis of this on the calculation of goniometric of system with the framework are used the formulas, derived for the calculation of the framework with inductive coupling.

In the case of the work of goniometric system in very wide frequency band, are applied the following methods of the realization of switching partial frequency bands: complete switching of goniometers (or their exchange) over partial ranges, switching the sections of search coil, application/use of diagrams of elongation and shortening for search coil and the application/use of intermediate untuned circuit.

Switching goniometers over partial frequency bands is the best method from purely electrical point of view. However, this method is bulky and very complicates mounting. Switching the sections of search coil of goniometer is applied very rarely, since it is extremely difficult to carry out this switching during fulfilling of all requirements in the relation to the symmetry of diagram and winding/coil, in the absence of spurious coupling and smallness of octant error of goniometer.

The use of a circuit of elongation and shortening of search coil is most widely used.

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For the calculation of the circuit of elongation, are directly used formulas (7.25) and (7.27) of present paragraph.

Let us give the calculation of the diagram of shortening, presented in Fig. 7.8a. In accordance with the already used previously method let us find equivalent emf and coupled impedance in the circuit of search coil:

$$E_{э}=E_1\frac{M}{L_0+L_1},\quad \Delta Z_r=\frac{M^2}{(L_0+L_1)^2}\,[R_0+R_1-j\omega(L_0+L_1)].$$

Let us desigaate the changed due to coupled impedances parameters of search coil $L'_к$ and $R'_к$. Then to further calculation is subject diagram Fig. 7.8b. It is completely similar to the diagram of the shortening of the framework. For determining edge stress and a resonance condition, we utilize formulas (7.20) and (7.21), replacing in them only the designations:

$$\omega_{1оэ}=\sqrt{\frac{L_к+L'_к}{L_кL'_кC}}=\frac{1}{\sqrt{L'C}},$$

$$U_{c1оэ}=E_2\frac{\omega L'_к}{R'_к+\frac{L'^2_к}{L^2_к}R_к}=E_1\frac{M}{L_0+L_1}\frac{\omega L'_к}{R'_к+\frac{L'^2_к}{L^2_к}R_к}, \quad (7.28)$$

$$p_э=h_c\frac{M}{L_0+L_1}\frac{\omega L'_к}{R'_к+\frac{L'^2_к}{L^2_к}R_к}. \quad (7.29)$$

In this formula are known all values, with the exception $L_к$, since they are determined by the calculation of that range on which shorting is absent. Thus, the calculation of the diagram of shortening is reduced to determination $L'_к$, then $L_к$ and to the subsequent checking of effectiveness from formula (7.29).

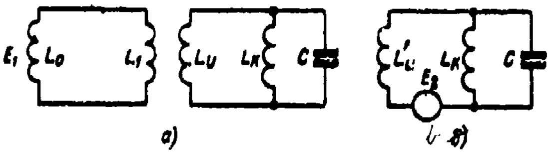

Fig. 7.8. Diagram of goniometer with shortening.

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Diagram with intermediate untuned circuit (Fig. 7.9) it is expedient to use, if it is necessary to decrease the reaction of frame duct to the tuned circuit. On effectiveness it can be favorable only with the very large number of frequency ranges. For its calculation let us find emf, induced in intermediate circuit:

$$E_{\text{пр}}=E_1\frac{M}{L_0+L_A}=E_1K_{L1}\frac{\sqrt{L_AL_0}}{L_A+L_0}.$$

The introduced into intermediate circuit resistor/resistance in this case we disregard, since it is not inclined, and therefore the effect of coupled impedance is small. The maximum of emf occurs when $L_{A\,\text{опт}}=L_0$ and it is equal to

$$E_{\text{пр макс}}=E_1\frac{K_{L1}}{2}\sqrt{\frac{L_A}{L_0}}.$$

To further calculation it is possible to use the formulas (7.24)-(7.27) of this paragraph, replacing in them L_0 on L_n. The transmission gain of system is determined by the formula

$$k_{\text{а макс}} = \frac{K_{L1}K_{L2}}{4\delta_2}\sqrt{\frac{\delta_2}{\delta_2 + K_{L2}^2\frac{L_{\text{св}}}{L_2}}}\frac{(2 - K_{L2}^2)\,\delta_2 + K_{L2}^2\delta_1\frac{L_{\text{св}}}{L_2}}{2\delta_2 + K_{L2}^2\delta_1\frac{L_{\text{св}}}{L_2}} \times$$

$$\times \sqrt{\frac{L_{\text{св}}}{L_0}}. \qquad (7.30)$$

The effectiveness of diagram with intermediate circuit is equal to the effectiveness of the diagram without intermediate circuit, multiplied by $\frac{K_{L1}}{2}$. In usual values $K_{L1} = 0{,}5 \div 0{,}7$ this corresponds to a decrease in the effectiveness 2-4 times.

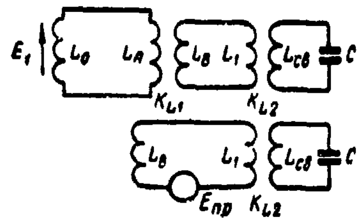

Fig. 7.9. Diagram of goniometer with unadjusted intermediate circuit.

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If loop antenna is referred from receiver and is connected with its input by long feeder (for example, with the aid of coaxial cable in ship radio direction finder), then in the given calculations instead of inductive reactance of framework ωL_0 and of emf. E_h will enter those who were converted due to those who are given below formulas (7.38)-(7.40) of resistor/resistance and emf of the framework. For a decrease in the manifestation of resonances sometimes at the end/lead of the feeder of receiver is connected supplementary effective resistance in parallel to the load of feeder or consecutively with it [7.13].

7.5. Calculation of effective height and input antenna resistance of system with a small separation of the vertical wire antennas (cosinusoidal directional characteristic).

The antenna system in question is either the rotary pair of antennas or n of motionless antennae arrange/located in the apex/vertexes of correct n-corner iron.

Let us examine the possible ways of fulfilling fundamental requirements on the calculation of by the antenna of system - obtaining minimum instrument errors and best sensitivity of radio direction finder.

In §§4.5 and 4.8 is given the calculation of the errors, produced by the different types of asymmetry. It was shown, that the value of errors and diffuseness of the minimum at direction finding is determined by the dissimilarity of amplitudes and phases of currents in separate vertical wire antennas, and also by the dissimilarity of the geometric dimensions of antennas and feeders. The small disagreement of the parameters of separate antennas and

feeders near the resonance frequency of antenna-feeder system produces the sharp inequality of amplitudes and phases of currents and, therefore, large errors and the diffuseness of the minimum. Experiment shows that with the usual accuracy of the production of antenna feeder system it is possible to ensure the acceptable level of errors, if the resonance frequencies of system differ from workers by 10-15o/o. Therefore resonance frequencies are selected so that they would lie/rest beyond range and would differ by the indicated value from one of the extreme frequencies of the range.

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From calculations it follows that the natural frequency ω_a of antenna feeder circuit is expedient to take lower than smallest frequencies of range $\omega_{мин}$, since in this case the effectiveness of radio direction finder changes in the range of frequencies less than when selecting the natural frequency the circuit ω_a, higher than greatest frequency of range $\omega_{макс}$. This is evident from Fig. 7.10 in which are constructed for certain special case two curves the dependences of the effectiveness of radio direction finder on the frequency when $\omega_a < \omega_{мин}$ and $\omega_a > \omega_{макс}$. ¶ In the first case the effectiveness changes over range less. Furthermore, on short waves when selecting natural frequency by the antenna of circuit higher than frequencies of operating range the inductance of the field coil

of goniometer can prove to be so insignificant that obtaining in the goniometer of uniform magnetic field becomes difficultly attained. In this case, also begins to be developed the magnetic distortion of goniometer as a result of the effect of the wires, which connect field coils with feeders.

In wide-range radio direction finder sometimes it is necessary to retain the resonance frequencies of antenna feeder system within the limits of the frequency band of the radio direction finder. In this case special importance has the careful adjustment of system on the sections of the frequencies, close to resonance.

In order to decrease the manifestation of asymmetry during approach/approximation to the resonance frequencies of antenna feeder system, one should as far as possible to lower wave antenna resistance, either introduce attenuation into feeders or select entry impedance of receiver, close to the wave impedance of feeder [7.7].

Earlier were obtained equivalent diagram (Fig. 3.33) and formula for the calculation of the current of search coil of goniometer (3.72).

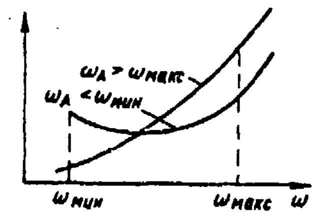

Fig. 7.10. Effectiveness curves of radio direction finder.

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If search coil of goniometer enters in the inclined grid circuit of the first tube of reception indicator and the capacitance/capacity of tuning is designated by C_c, that for a grid voltage of the first tube we will obtain

$$U_{c\ рез} = I_{н\ макс} \frac{1}{\omega C_c},$$

whence the effectiveness of goniometric system with n antennae will be

$$p_0 = \frac{U_{c\,\text{вых}}}{E} =$$

$$= \operatorname{Mod}\left[\frac{\omega M_э H_д \sqrt{\frac{N}{2}}\,\frac{1}{\omega C_c}}{(Z_{аф} + j\omega L_{пэ})\left(j\omega L_и + Z_и + \frac{\omega^2 M_э^2}{Z_{аф} + j\omega L_{пэ}}\right)}\right], \quad (7.31)$$

where N = n/2 is a number of pairs of antennas;

H_0 - effective height vapors of diametrically opposite antennas;

$L_{пэ} = L_п \frac{N}{2}$ - equivalent according to diagram inductance in antenna circuit;

$L_п$ is inductance of the field coil of goniometer;

$L_и$ - the inductance of search coil of goniometer;

$Z_и$ - the load impedance of search coil;

$M_э = K\sqrt{L_{пэ} L_и}$ is mutual inductance according to the diagram:

$$M_э = M_{макс}\sqrt{\frac{N}{2}};$$

$M_{макс}=K\sqrt{L_п L_н}$ is maximum mutual inductance between the coils of goniometer;

K - maximum coupling coefficient in goniometer and coupling coefficient in equivalent diagram;

$$Z_{аф}=2Z_{а0}+2\sum_{m=1}^{n}Z_{ом\,а}\cos\frac{2\pi m}{n} \qquad (7.32)$$

impedance of the pair of antennas, referred to current in the field coil of goniometer (§3.10).

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In the case of the rotary pair of antennas $M_{макс}$, K, $L_п$ and $L_н$ they are related to the input transformer of receiving indicator.

The problem concerning application/use in the diagram of intermediate untuned circuit must be solved, combining requirements of sensitivity and one-handed tuning.

For the calculation of effectiveness p_0 by the antenna of system it is required to determine the parameters, entering the

formula (7.31).

Effect of the number of antennas.

For an increase in the effectiveness at the base altitude of single vertical wire antenna, it is possible to increase the number of antennas and the separation between them. The value of separation is limited to the maximum error of separation, which is observed on smallest wave of operating range. The permissible separation with the assigned error increases with an increase in the number of antennas. Thus, an increase in the number of antennas is led to an increase in the effectiveness not only because of the number of antennas, but also because of an increase in the permissible separation.

For the quantitative determination of the effect of an increase in the number of antennas by effectiveness, let us turn to formula (7.31).

Calculations show that the total resistance of the pair of antennas $Z_{a\phi}$ little depends on the number of antennas. Therefore under the condition of the preservation/retention/maintaining of the resonance frequency of the pair of antennas, we have

$$Z_{a\phi} + i\omega L_{na} \approx \text{const.}$$

If coupling coefficient in goniometer does not change, then we come to the conclusion that L_{na} and M_{a} they do not depend on the number of antennas. From formula (7.31) it follows that in this case the effectiveness by the antenna of system p_{a} is proportional to square root from the number of antennas. Effectiveness changes also due to a change in the effective effective height H_0, which affects separation 2b, different for the different number of antennas (see Table 4.1).

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Calculation of effective height and entry impedance of antenna feeder system from the pair of antennas with the direct connection of antennas to feeders.

We give the necessary for further calculations formulas from the theory of long lines [3.1, 3.4].

Let voltage E_{a} be connected through resistor/resistance $\dot{Z}_{a}$ to the input of feeder. Let us designate $m_{\phi}, \beta_{\phi}, \gamma_{\phi}$ - the constant

of phase displacement, attenuation and propagation constant of feeder, moreover

$$\gamma_\Phi=\beta_\Phi+jm_\Phi;\ m_\Phi=m\sqrt{\varepsilon_\Phi}\ ;\ m=\frac{2\pi}{\lambda},$$

ε_Φ - the dielectric constant of insulation of feeder;

ρ_Φ - wave impedance;

l_Φ - the length of feeder;

$Z_a=\rho_\Phi \operatorname{cth}\theta_a.$

Then output resistance at the end/lead of the feeder will be

$$Z_{\text{вых}}=\rho_\Phi \operatorname{cth}(\gamma_\Phi l_\Phi+\theta_a)=\rho_\Phi\frac{1+\operatorname{th}\theta_a \operatorname{th}\gamma_\Phi l_\Phi}{\operatorname{th}\gamma_\Phi l_\Phi+\operatorname{th}\theta_a}=$$
$$=\rho_\Phi\frac{Z_a \operatorname{cth}\gamma_\Phi l_\Phi+\rho_\Phi}{Z_a+\rho_\Phi \operatorname{cth}\gamma_\Phi l_\Phi}=R_{\text{вых}}+jX_{\text{вых}}. \qquad (7.33)$$

Voltage on the end/lead of the feeder is expressed by the formula

$$E_{\text{вых}}=\frac{E_a}{\operatorname{ch}\theta_a}\operatorname{ch}(\gamma_\Phi l_\Phi+\theta_a)=E_a\left(\operatorname{ch}\gamma_\Phi l_\Phi+\frac{\rho_\Phi}{Z_a}\operatorname{sh}\gamma_\Phi l_\Phi\right). \qquad (7.34)$$

We convert (7.33) and (7.34):

$$E_{\text{вых}} = \frac{\sqrt{2}\rho_\Phi E_a e^{-jm_\Phi l_\Phi}}{\sqrt{[(z_a^2+\rho_\Phi^2)\operatorname{ch}2\beta_\Phi l_\Phi + 2R_a\rho_\Phi \operatorname{sh}2\beta_\Phi l_\Phi] + [(\rho_\Phi^2 - z_a^2)\cos 2m_\Phi l_\Phi - 2X_a\rho_\Phi \sin 2m_\Phi l_\Phi]}}, \tag{7.35}$$

$$R_{\text{вых}} = \frac{\rho_\Phi [(\rho_\Phi^2 + z_a^2)\operatorname{sh}2\beta_\Phi l_\Phi + 2R_a\rho_\Phi \operatorname{ch}2\beta_\Phi l_\Phi]}{[(z_a^2+\rho_\Phi^2)\operatorname{ch}2\beta_\Phi l_\Phi + 2R_a\rho_\Phi \operatorname{sh}2\beta_\Phi l_\Phi] + [(\rho_\Phi^2 - z_a^2)\cos 2m_\Phi l_\Phi - 2X_a\rho_\Phi \sin 2m_\Phi l_\Phi]}, \tag{7.36}$$

$$X_{\text{вых}} = \frac{\rho_\Phi [(\rho_\Phi^2 - z_a^2)\sin 2m_\Phi l_\Phi + 2X_a\rho_\Phi \cos 2m_\Phi l_\Phi]}{[(z_a^2+\rho_\Phi^2)\operatorname{ch}2\beta_\Phi l_\Phi + 2R_a\rho_\Phi \operatorname{sh}2\beta_\Phi l_\Phi] + [(\rho_\Phi^2 - z_a^2)\cos 2m_\Phi l_\Phi - 2X_a\rho_\Phi \sin 2m_\Phi l_\Phi]}. \tag{7.37}$$

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Here

$$z_a = \sqrt{R_a^2 + X_a^2}.$$

When

$$Z_a = R_a = \rho_\Phi \quad \text{and} \quad X_a = 0,$$

that

$$E_{\text{вых}}=\frac{E_{\text{a}}e^{-jm_{\phi}l_{\phi}}}{\sqrt{\text{ch}\,2\beta_{\phi}l_{\phi}+\text{sh}\,2\beta_{\phi}l_{\phi}}}=E_{\text{a}}e^{-(\beta_{\phi}+jm_{\phi})l_{\phi}},$$

$$R_{\text{вых}}=\rho_{\phi},\ X_{\text{вых}}=0.$$

The field coil of goniometer it is possible to inclale/connect consecutively or parallel in the feeder of antenna. Let us examine these cases separately.

Figures 7.11 gives schematic and equivalent diagrams for calculation by the antenna of system with the series connection of the field coil of goniometer into feeder from antennas (U-shaped system). In figure $Z_{a1}=R_{a1}+jX_{a1}$ and, $Z_{a2}=R_{a2}+jX_{a2}$ are impedances of separate antennas (its own and introduced from the adjacent antennas), referred to the current antinode in antenna; $E_{a1}=Eh_{e}e^{-jmb}$ and $E_{a2}=Eh_{e}e^{jmb}$ - emf, induced in antennas.

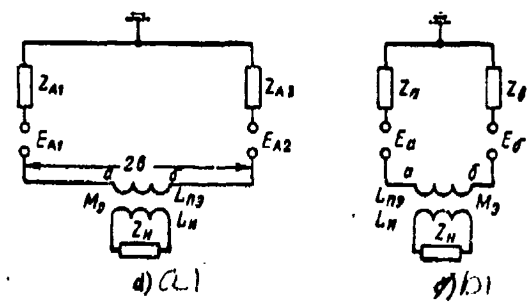

Fig. 7.11. Sequential switching circuits of field coil (J-shaped system): a) schematic diagram; b) equivalent diagram.

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Let us convert resistor/resistance Z_{a1} and emf E_{a1} of left vibrator to the point a of the connection of field coil. Disregarding the attenuation of feeders, we will obtain the formulas (7.35)-(7.37):

$$E_a=\frac{\rho_\Phi E h_\partial e^{-jmb}}{\sqrt{\rho_\Phi^2\cos^2 m_\Phi l_\Phi+(R_{a1}^2+X_{a1}^2)\sin^2 m_\Phi l_\Phi-X_{a1}\rho_\Phi\sin 2m_\Phi l_\Phi}}, \tag{7.38}$$

$$R_a=\frac{R_{a1}\rho_\Phi^2}{\rho_\Phi^2\cos^2 m_\Phi l_\Phi+(R_{a1}^2+X_{a1}^2)\sin^2 m_\Phi l_\Phi-X_{a1}\rho_\Phi\sin 2m_\Phi l_\Phi}, \tag{7.39}$$

$$X_a=\frac{\rho_\Phi\left[(\rho_\Phi^2-R_{a1}^2-X_{a1}^2)\sin m_\Phi l_\Phi\cos m_\Phi l_\Phi+X_{a1}\rho_\Phi-2X_{a1}\rho_\Phi\sin^2 m_\Phi l_\Phi\right]}{\rho_\Phi^2\cos^2 m_\Phi l_\Phi+(R_{a1}^2+X_{a1}^2)\sin^2 m_\Phi l_\Phi-X_{a1}\rho_\Phi\sin 2m_\Phi l_\Phi}. \tag{7.40}$$

For the right vibrator of expression for $R_б$ and $X_б$ remain the same, for voltage $\dot{E}_б$ (at point b) in numerator (7.38) instead of e^{-jmb} one should write e^{jmb}.

the voltage at points ab with off field coil will be

$$E_{aб}=E_{a\Phi}=E_a-E_б. \tag{7.41}$$

The affective height of the antenna feeder system of the pair of antennas from (7.38) and (7.41) when $Z_{a1}\doteq Z_{a2}=Z_A$ is obtained equal to

$$H_e = \frac{E_{a\phi}}{E} =$$

$$= \frac{2 \sin mb\, h_e \rho_\phi}{\sqrt{\rho_\Phi^2 \cos^2 m_\phi l_\phi + (R_a^2 + X_a^2) \sin^2 m_\phi l_\phi - X_a \rho_\phi \sin 2m_\phi l_\phi}}. \quad (7.42)$$

Entry impedaice of the antenna feeder system of the pair of antennas, in reference to current in the field coil of goniometer, will be

$$Z_{a\delta} = Z_{a\phi} = Z_a + Z_\delta.$$

With the complete symmetry of system $Z_a = Z_\delta$ and therefore

$$Z_{a\phi} = 2Z_a = 2(R_a + jX_a). \quad (7.43)$$

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Formulas (7.42), (7.43), (7.39) and (7.40) they serve for the calculation of effective height and resistor/resistance of antenna feeder system when it is necessary to consider the resistor/resistances of vibrators R_a. If it is possible to disregard R_a in comparison with X_a and to count that

$$X_a = -\rho_B \operatorname{ctg} ml_B,\; h_e = \frac{1}{m} \operatorname{tg} \frac{ml_B}{2},$$

where ρ_B, l_B - wave impedance and the length of vibrator, then

formulae (7.39), (7.42), (7.40) and (7.43) can be simplified

$$R_a = 0$$

$$H_o = \frac{2\sin mb\,\mathrm{tg}\,\frac{ml_n}{2}}{m\left(\cos m_\Phi l_\Phi + \frac{\rho_n}{\rho_\Phi}\sin m_\Phi l_\Phi\,\mathrm{ctg}\,ml_n\right)}, \quad (7.44)$$

$$X_{a,б} = \frac{\rho_\Phi(\rho_\Phi \sin m_\Phi l_\Phi - \rho_n\,\mathrm{ctg}\,ml_n \cos m_\Phi l_\Phi)}{\rho_\Phi \cos m_\Phi l_\Phi + \rho_n\,\mathrm{ctg}\,ml_n \sin m_\Phi l_\Phi} =$$

$$= \rho_\Phi \frac{1 - \frac{\rho_n}{\rho_\Phi}\,\mathrm{ctg}\,ml_n\,\mathrm{ctg}\,m_\Phi l_\Phi}{\mathrm{ctg}\,m_\Phi l_\Phi + \frac{\rho_n}{\rho_\Phi}\,\mathrm{ctg}\,ml_n}, \quad (7.45)$$

$$Z_{a\Phi} = j2X_{a,б}. \quad (7.46)$$

Figures 7.12 gives schematic and equivalent diagrams for calculation by the antenna of system with the parallel connection of the field coil of goniometer into feeder from antennas (H-shaped system).

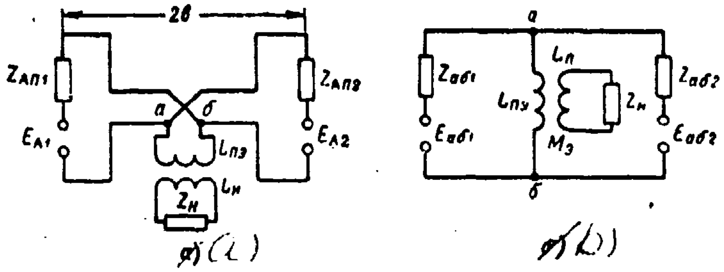

Fig. 7.12. Circuits of parallel connection of field coil (H-shaped system): a) schematic diagram; b) equivalent diagram.

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In Fig. 7.12 designations the same as in Fig. 7.11. Voltages $E_{аб1}$ and $E_{аб2}$ and resistor/resistances $Z_{аб1}$ and $Z_{аб2}$ from each antenna at

pointsab of the connection of field coil of goniometer are designed from the same formulas (7.38), (7.39) and (7.40), that also for the case of the series connection of field coil.

The total voltage from both of antennas on the terminal/grippers of field coil with its disconnection/cutoff in accordance with Fig. 7.12b will be

$$E_{\text{аф}}=\frac{E_{\text{аб}2}Z_{\text{аб}1}}{Z_{\text{аб}1}+Z_{\text{аб}2}}-\frac{E_{\text{аб}1}Z_{\text{аб}2}}{Z_{\text{аб}1}+Z_{\text{аб}2}}.$$

Entry impedance of antenna feeder system on terminal/grippersab is determined by the expression

$$Z_{\text{аф}}=\frac{Z_{\text{аб}1}Z_{\text{аб}2}}{Z_{\text{аб}1}+Z_{\text{аб}2}}.$$

When diagram is symmetrical, then

$$Z_{\text{аб}1}=Z_{\text{аб}2}=Z_{\text{аб}},\ Z_{\text{аф}}=\frac{Z_{\text{аб}}}{2},\quad E_{\text{аф}}=\frac{E_{\text{аб}2}-E_{\text{аб}1}}{2}.$$

Expressions for H_0 and $Z_{\text{аф}}$ in the case of parallel connection of field coil of goniometer into the feeder of antennas will be

$$H_{\text{ф}}=\frac{\sin mb\, h_{\text{д}}\rho_{\text{ф}}}{\sqrt{\rho_{\text{ф}}^2\cos^2 m_{\text{ф}}l_{\text{ф}}+(R_{\text{а}}^2+X_{\text{а}}^2)\sin^2 m_{\text{ф}}l_{\text{ф}}-X_{\text{а}}\rho_{\text{ф}}\sin 2m_{\text{ф}}l_{\text{ф}}}}, \tag{7.47}$$

$$Z_{\text{аф}}=\frac{Z_{\text{аб}}}{2}=\frac{1}{2}R_{\text{аб}}+j\frac{1}{2}X_{\text{аб}}, \tag{7.48}$$

where R_{a6} and X_{a6} they are designed from formulas (7.39) and (7.40).

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If it is possible to disregard resistor/resistance R_a in comparison with X_a and to count that in this case

$$X_{a_\circ} \approx -\rho_a \operatorname{ctg} ml_a, \; h_e = \frac{2}{m} \operatorname{tg} \frac{ml_a}{2},$$

we will obtain the calculation formulas

$$H_{0}=\frac{2\sin mb\,\mathrm{tg}\,\frac{ml_{\mathrm{n}}}{2}}{m\left(\cos m_{\Phi}l_{\Phi}+\frac{\rho_{\mathrm{n}}}{\rho_{\Phi}}\sin m_{\Phi}l_{\Phi}\,\mathrm{ctg}\,ml_{\mathrm{n}}\right)}, \quad (7.49)$$

$$X_{\mathrm{n}\Phi}=\frac{1}{2}\,\frac{\rho_{\Phi}\left(1-\frac{\rho_{\mathrm{n}}}{\rho_{\Phi}}\,\mathrm{ctg}\,m_{\Phi}l_{\Phi}\,\mathrm{ctg}\,ml_{\mathrm{n}}\right)}{\mathrm{ctg}\,m_{\Phi}l_{\Phi}+\frac{\rho_{\mathrm{n}}}{\rho_{\Phi}}\,\mathrm{ctg}\,ml_{\mathrm{n}}};\ R_{\mathrm{n}\Phi}=0. \quad (7.50)$$

When $l_{\mathrm{n}}\ll\lambda$ and $b\ll\lambda$, formula (7.44), (7.45), (7.49) and (7.50) they are considerably simplified, since it is possible to count

$$\sin mb=mb,\ \sin\frac{ml_{\mathrm{n}}}{2}=\frac{ml_{\mathrm{n}}}{2},\ \cos m_{\Phi}l_{\Phi}=1,$$

$$\mathrm{ctg}\,m_{\Phi}l_{\Phi}=\frac{1}{m_{\Phi}l_{\Phi}},\ \ \mathrm{ctg}\,ml_{\mathrm{n}}=\frac{1}{ml_{\mathrm{n}}}$$

and the product of sines equal to zero.

Then for the series connection of the field coil

$$H_0 = \frac{2\pi l_{\text{в}} b}{\lambda\left(1+\frac{C_{\phi} l_{\phi}}{C_{\text{в}} l_{\text{в}}}\right)} = -\frac{2\pi l_{\text{в}} b}{\lambda\left(1+\frac{C_{\phi\text{п}}}{C_{\text{вп}}}\right)}, \tag{7.51}$$

$$X_{\text{а}\phi} = \frac{2}{\rho_{\text{в}}}\,\frac{\rho_{\phi}\sin ml_{\text{в}}\sin m_{\phi}l_{\phi} - \rho_{\text{в}}\cos ml_{\text{в}}\cos m_{\phi}l_{\phi}}{\frac{1}{\rho_{\text{в}}}\cos m_{\phi}l_{\phi}\sin ml_{\text{в}} + \frac{1}{\rho_{\phi}}\cos ml_{\text{в}}\sin m_{\phi}l_{\phi}} =$$

$$= -\frac{2}{\frac{ml_{\text{в}}}{\rho_{\text{в}}}+\frac{m_{\phi}l_{\phi}}{\rho_{\phi}}} = -\frac{2}{\omega\,(C_{\phi\text{п}}+C_{\text{вп}})}. \tag{7.52}$$

For the parallel connection of the field coil

$$H_0 = \frac{2\pi l_{\text{в}} b}{\lambda\left(1+\frac{C_{\phi\text{п}}}{C_{\text{вп}}}\right)}, \tag{7.53}$$

$$X_{\text{а}\phi} = -\frac{1}{2\omega\,(C_{\phi\text{п}}+C_{\text{вп}})}, \tag{7.54}$$

where $C_{\text{в}}, C_{\phi}$ are linear capacitance/capacities of vibrator and feeder on 1 m of length;

$C_{\text{вп}}=C_{\text{в}}l_{\text{в}},\ C_{\phi\text{п}}=C_{\phi}l_{\phi}$ are complete antenna capacities and feeder.

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From formulas (7.51) and (7.53) is visible the importance of a decrease in the relation $\frac{C_{\phi n}}{C_{sn}}$ for an increase in the effective height of system. For the possibility of the calculation of effectiveness according to formula (7.31) we will examine the questions of selection n and of calculation H_0 and $Z_{a\phi}$. The parameters L_n, Z_n, K and C_c are establish/installed during the calculation of input circuit of reception indicator. Procedure of calculation L_{n3} is clarified below.

7.6. Calculation of H-shaped system.

Fundamental and the equivalent diagrams of the pair of antennas are given in Fig. 7.12.

For calculation is assigned the frequency band of the radio direction finder $f_{мин}$—$f_{макс}$.

Let us calculate first of all antenna-dipole. The length of halfdipole $l_в$ we will select on the basis of the fact that in vertical radiation pattern will not being gaps. Must be made the

condition

$$l_{в} < 0{,}625\lambda_{мин}.$$

The diameter of dipole 2a it is desirable to take as possible large, so that by an increase in the diameter is improved the agreement of dipole with feeder and increases H_0. We will calculate $C_{в}, \rho_{в}, R_{\Sigma}, R_{а}, X_{а}, \beta_e$ of dipole in the range of frequencies $f_{мин} - f_{макс}$, and also h_e for this same frequency band. Considerations on selection of separation 2b are presented earlier. On the basis of construction, we determine $2l_{\Phi}$. It is selected or we design feeder. It is desirable so that ρ_{Φ} was of one order or is more than $\rho_{в}$. We design from formulas (7.47)-(7.50) $Z_{аб}$ and $Z_{в\Phi} = 0{,}5\ Z_{аб}$, and also

$$H_0 = \left| \frac{E_{аб1} - E_{аб2}}{2E} \right|.$$

The inductance of field coil $L_{п} = \frac{L_{пэ}}{n/2}$ it is selected so that the natural frequency of the antenna feeder circuit of the pair of antennas, which consists of the pair of antennas, feeder and the equivalent field coil of goniometer $L_{пэ}$ (see Fig. 3.33), it was located outside working frequency band $f_{мин} - f_{макс}$.

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Figures 7.13 depicts the dependence of the reactance of antenna feeder circuit (taking into account the resistor/resistances,

introduced from other antennas), undertaken with opposite sign $(-X_{a\phi})$ on frequency. Two Possible straight lines for $L_{n\vartheta}=L_n/\frac{N}{2}$ are given in this same figure. The natural frequencies of antenna feeder circuit are determined by the points of intersection of curve $-X_{a\phi}$ and straight lines $\omega L_{n\vartheta}$; they are selected so that they differ by 10-15o/o from the extreme frequencies of the range ($f_{мин}$ and $f_{макс}$).

As shown earlier, it is profitable to stop at straight line, giving intersection from curve at frequency less $\omega_{мин}$. Straight line $\omega L_{n\vartheta}$ determines the inductance L_n of field coil, for example

$$L_n=\frac{AB}{\omega_{мин}}\frac{1}{N/2},$$

if the self-capacitance of field coil can be disregarded.

In the general case the resistor/resistance of the field coil of goniometer X_n is parallel connection of inductance L_n and of coil capacitance C_s. Therefore, if it is not possible to disregard capacitance/capacity C_s, then

$$X_n=\frac{\omega L_n}{1-L_nC_s\omega^2}$$

and in the preceding/previous example

$$L_n=\frac{1}{N/2}\frac{AB}{\omega_{мин}(1+AB\omega^2_{мин}C_s)}. \qquad (7.55)$$

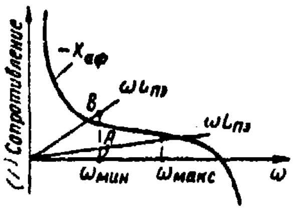

Fig. 7.13. Selection of inductance of field coil of goniometer.

Key: (1). Resistor/resistance.

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In §4.8 are derived the dependences of the errors of spaced-antenna direction finder on the different reasons for asymmetry of antenna feeder system and is made the conclusion about the need for designing this system, so that its resonance frequencies lie/rest outsidethe working frequency band of the radio direction finder.

Let us examine, which limitations this requirement places on the

selection of size/dimensions of the antenna of system.

Let us dismantle/select two cases:

1) the reception/procedure of the vertically polarized electric field;

2) the reception/procedure of the horizontally polarized electric field.

Reception/procedure of the vertically polarized field.

Considering that in the center of system the phase of electric field is equal to zero, we have for emf in dipoles (see Fig. 7.12a)

$$\dot{E}_{a_1} = \dot{E}h_a e^{-j\varphi},$$
$$\dot{E}_{a_2} = \dot{E}h_a e^{j\varphi},$$

where

$$\varphi = \frac{2\pi}{\lambda} b \cos\theta \cos\beta.$$

Decompose emf E_{a_1} and E_{a_2} into phase $E_{\text{ф}}$ and nonphase $E_{\text{нф}}$ terms:

$$E_{a_1} = \dot{E}h_a(\cos\varphi - j\sin\varphi) = E_{1\text{ф}} + jE_{1\text{нф}},$$
$$E_{a_2} = \dot{E}h_a(\cos\varphi + j\sin\varphi) = E_{2\text{ф}} + jE_{2\text{нф}}$$

or

$$E_{1\text{ф}} = Eh_e \cos\left(\frac{2\pi}{\lambda} b \cos\theta \cos\beta\right),$$
$$E_{1\text{нф}} = -Eh_e \sin\left(\frac{2\pi}{\lambda} b \cos\theta \cos\beta\right),$$
$$E_{2\text{ф}} = E_{1\text{ф}} \quad E_{2\text{нф}} = -E_{1\text{нф}}.$$

Phase the components of emf determine one-act current, nonphase - the push-pull current, passing through the field coil of goniometer.

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It is obvious that phase the components of emf in balanced network will not create current in field coil, i.e., the voltages at points a and b will be equal to each other (Fig. 7.14a). Equivalent diagram for the single-cycle current of system will obtain form as in Fig. 7.14b, with short circuit instead of the coil.

Nonphase the components of emf of right and the left of dipoles store/add up, and they will create in field current coil. The equivalent diagram of system for a push-pull current will take the form, depicted on Fig. 7.15, where $X_{\text{пэ}}$ is the reactance of the field coil of goniometer. In Fig. 7.15 are designated instantaneous values

emf.

Resonance condition for a single-cycle current from Fig. 7.14b will be

$$X_{\text{в}}+\rho_{\text{ф}}\operatorname{tg} m_{\text{ф}} l_{\text{ф}}=0$$

or, considering that

$$X_{\text{в}}=-\rho_{\text{в}}\operatorname{ctg} m l_{\text{в}},$$

we have

$$\rho_{\text{ф}}\operatorname{tg} m_{\text{ф}} l_{\text{ф}}-\rho_{\text{в}}\operatorname{ctg} m l_{\text{в}}=0,$$

whence

$$1-\frac{\rho_{\text{в}}}{\rho_{\text{ф}}}\operatorname{ctg} m l_{\text{в}}\operatorname{ctg} m_{\text{ф}} l_{\text{ф}}=0, \qquad (7.56)$$

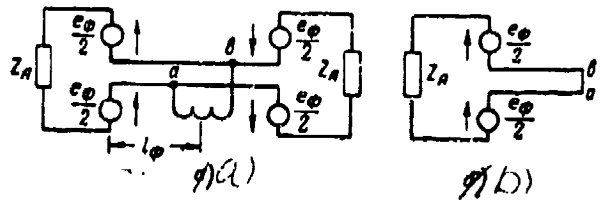

Fig. 7.14. Equivalent diagrams for phase currents.

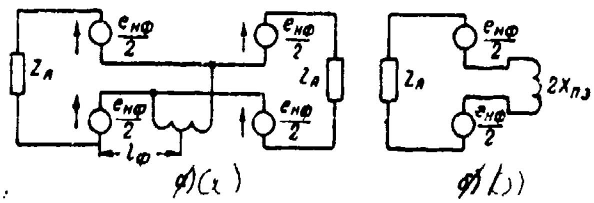

Fig. 7.15. Equivalent diagram for nonphase currents.

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In order to avoid resonance for a single-cycle current, it is

necessary to satisfy in working frequency band the condition

$$l_{\text{в}} \neq \frac{1}{m} \operatorname{arcctg}\left(\frac{\rho_{\Phi}}{\rho_{\text{в}}} \operatorname{tg} m_{\Phi} l_{\Phi}\right). \tag{7.57}$$

Let this resonance occur for frequency, to 10-15o/o of larger maximum frequency of range. Let us designate m and m_{Φ} for frequency by 10-15o/o higher than maximum by m_0 and $m_{\Phi 0}$. Then supplementary limitation for the selection of the length of halfdipole $l_{\text{в}}$ at the selected length of feeder l_{Φ} will be from (7.57):

$$l_{\text{в}} < \frac{1}{m_0} \operatorname{arcctg}\left(\frac{\rho_{\Phi}}{\rho_{\text{в}}} \operatorname{tg} m_{\Phi 0} l_{\Phi}\right). \tag{7.57'}$$

During the calculation of resonance frequencies for a push-pull current, we proceed from Fig. 7.15b in which is reject/thrown the half of antenna feeder system and therefore $X_{\text{пэ}}$ is replaced by $2X_{\text{пэ}}$.

the condition of the absence of resonance for a push-pull current in the field coil of goniometer, as this follows from equivalent diagram in Fig. 7.15b, coincides on condition for the selection of the inductance of field coil. Thus, during the correct selection of the inductance of field coil this condition automatically is satisfied.

Reception/procedure of the horizontally polarized field.

In by normally working H-type antenna to system the horizontal component of electric field is not produced currents in the field coils of goniometer due to the symmetry of system.

If in system appears asymmetry (for example, inequality of resistances of one of the half-dipoles relative to the resistor/resistances of other half-dipoles), then the currents, which take place in the halves of field coil from horizontal field, are not counterbalanced. Are observed the reception/procedure of horizontal field and polarizational errors. So that asymmetry of system strongly does not manifest itself, the reactance of system in working frequency band for phase and nonphase voltages of horizontal electric field must not be equal to zero.

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for the horizontally polarized field, inducing emf in feeders, the feeders must be considered as single-wire, and lower and upper halfdipoles - as connected in parallel between the end/leads of single-wire feeders and the earth/ground. Equivalent diagram for this after becoming it is depicted on Fig. 7.16a.

Let ρ'_Φ - the wave impedance of the feeders, considered as bunched conductors; it depend on outer diameter of the shells of the feeder

$$\rho'_\Phi = \frac{3333}{C'_\Phi}, \text{ ohm}$$

where C'_Φ is a linear capacitance/capacity earth referenced both feeders, n p.

The halves of the field coils of goniometer as are included in parallel. Since currents in these halves are directed to reverse sides, the resulting inductance is close to zero and reactance them can be disregarded.

In figure C_k - the capacitance/capacity of coil earth referenced;

$\frac{X_A}{4}$ - shunt resistance of upper and lower halfdipoles.

Let, as earlier, field at the center of system has zero phase.

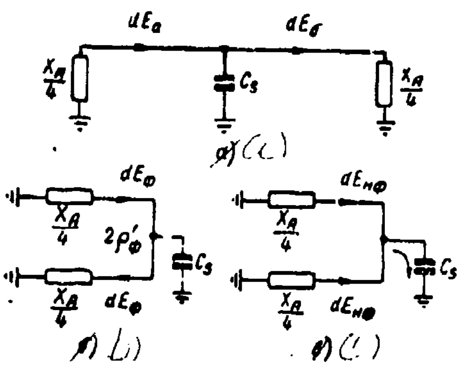

Fig. 7.16. Equivalent diagrams for reception/procedure of horizontally polarized field.

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The electromotive forces, induced in the cell/elements of right and by the left of halves, equidistant on x from center, can be expressed thus:

$$\left.\begin{aligned} dE_a &= dE e^{-jmx} = dE_{\Phi} - jdE_{\text{н}\Phi}, \\ dE_b &= dE e^{jmx} = dE_{\Phi} + jdE_{\text{н}\Phi}. \end{aligned}\right\} \qquad (7.58)$$

Phase the components of emf dE_{Φ} are directed both in one side, nonphase $dE_{\text{н}\Phi}$ - into different.

If we accumulate the beams of feeders into two-wire circuit, then equivalent diagrams for phase and nonphase stress component will take the form, depicted on Fig. 7.16b and 7.16c respectively.

Phase current occurs besides C_x therefore in Fig. 7.16b connection with C_x shown by dotted line.

The condition of equality zero of the reactive component of resistor/resistance for a phase current (Fig. 7.16b) will be

$$\frac{X_{\text{в}}}{2} + 2\rho'_{\Phi} \operatorname{tg} ml_{\Phi} = 0. \qquad (7.59)$$

For the feeders as of bunched conductors m = 2π/λ - propagation

constant.

After replacing $\dot{X}_в = -\rho_в \operatorname{ctg} ml_в$, we will obtain from (7.59)

$$l_в = \frac{1}{m} \operatorname{arcctg}\left(\frac{4\rho'_\Phi}{\rho_в} \operatorname{tg} ml_\Phi\right). \qquad (7.59')$$

Condition of equality zero of reactance for a nonphase current (Fig. 7.16c):

$$\frac{\dot{X}_в}{4} - \rho'_\Phi \operatorname{ctg} ml'_\Phi = 0. \qquad (7.60)$$

The equivalent length of feeder l'_Φ is obtained as a result of the account of terminal capacitance/capacity $C_э$ and is determined by the condition

$$\rho'_\Phi \operatorname{ctg} ml'_\Phi = \frac{\operatorname{ctg} ml_\Phi \frac{2}{\omega C_э} - \rho'_\Phi}{\operatorname{ctg} ml_\Phi + \frac{2}{\omega C_э \rho'_\Phi}} = \frac{2\rho'_\Phi \operatorname{ctg} ml_\Phi - \rho'^2_\Phi \omega C_э}{\rho'_\Phi \omega C_э \operatorname{ctg} ml_\Phi + 2}$$

or approximately

$$l'_\Phi \approx l_\Phi + \frac{C_э}{2C'_\Phi}. \qquad (7.61)$$

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Substituting in (7.60) the value

$$X_в = -\rho_в \operatorname{ctg} ml_в,$$

we will obtain for the resonance frequency

$$l_в = \frac{1}{m} \operatorname{arcctg}\left(-4\frac{\rho'_\Phi}{\rho_в} \operatorname{ctg} ml'_\Phi\right). \qquad (7.62)$$

where l'_Φ it is determined by equality (7.61).

In order to avoid in the working frequency band of the radio direction finder of the resonance frequencies of the antenna feeder system for the reception/procedure of horizontal electric field, expression (7.59'), and (7.62) for $l_\text{п}$ they must be made at frequencies, which lie higher than working frequency band of the radio direction finder. Let us designate m_0 value for m at the frequency, greater than the maximum frequency of range by 10-15o/o. Then must be made the conditions:

$$\left.\begin{aligned} l_\text{в} &\leqslant \frac{1}{m_0} \operatorname{arcctg}\left[4\,\frac{\rho'_\Phi}{\rho_0}\operatorname{tg} m_0 l_\Phi\right], \\ l_\text{в} &\leqslant \frac{1}{m_0} \operatorname{arcctg}\left[-4\,\frac{\rho'_\Phi}{\rho_0}\operatorname{ctg} m_0 l'_\Phi\right]. \end{aligned}\right\} \quad (7.63)$$

Calculation of effectiveness.

The optimum coupling coefficient field and search coils has smallest value at the extreme frequency of the working frequency band

of the radio direction finder, the closest to natural frequency antenna feeder circuit. It is designed from the formula

$$K_0 = \sqrt{1 + \frac{X_A}{2X_{a0}}}\sqrt{\frac{Q_A}{Q_K}},$$

where Q_A is an energy factor of antenna circuit;

Q_K - the energy factor of input circuit of receiver.

Let us assume that in goniometer from the viewpoint of the permissible octant errors it is possible to take coupling coefficient K_r.

It is selected for K of goniometer smaller of the values K_0 and K_r. The duct of search coil we design, on the basis of assigned adjustable capacitor, overlaps and selected diagram of input circuit.

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We check coupled impedances from by the antenna of circuit into the grid circuit of the first tube. If the introduced reactance cannot be compensated for in the range of frequencies, we decrease K or is introduced into diagram intermediate untuned circuit.

After the determination of all cell/elements of the input part of the radio direction finder, we design the effectiveness of the

antenna feeder system of the pair of antennas and entire radio direction finder.

7.7. Calculation of U-shaped system.

The separation between antennas 2b is selected from the considerations of the permissible error of separation, the height of the vertical wire antenna l_B - on the basis of vertical directional characteristic.

Separate vertical wire antenna we design just as in H-shaped system, only one should consider that for the asymmetric vibrator of U-shaped system ρ_B, R_Σ and h_e they will be two times less than for the symmetrical vibrator of H-shaped system.

Coupled impedances due to the communication/connection between antennas are determined from curves or from formulas, given in the courses of antennas [3.1, 3.4]. In Fig. 7.17 are designated the distances between antennas, which determine these resistor/resistances.

Further we determine $Z_A = R_A + jX_A$ - complete input

antenna resistance. Is selected the type of feeder, i.e., we find that ρ_ϕ. The length of feeder l_ϕ we take, on the basis of the construction of system (with the buried feeders, with the feeders, placed under wire gauze, and so forth).

We design $Z_a = Z_б = R_a + jX_a$ from formulas (7.39) and (7.40) or (7.45), (7.54). Impedance

$$Z_{a\phi} = Z_a + Z_б = 2Z_a.$$

The effective height of the antenna feeder system of the pair of antennas H_0 we design from formulas (7.42), (7.44) or (7.51).

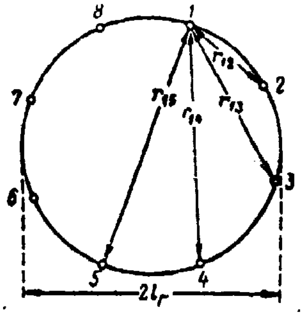

Fig. 7.17. On the calculation of mutual impedances.

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The inductance of field coil it is selected so that the natural frequency of the antenna feeder circuit of the pair of antennas together with the field coil of goniometer (Fig. 7.13) was located outside working frequency band $\omega_{мин} - \omega_{макс}$.

Let us restrict the effect of asymmetry of systems by means of this selection of the size/dimensions of system so that would not be developed the resonance phenomena.

In §6.3 it is shown, that for the limitation of current in the shell of the feeder, created by the reception/procedure of the horizontal component of electric field, must be made the condition $Z_з \ll Z_{аб}$,
where $Z_з$ - earth resistance of the shell of feeder and $Z_{аб}$ - the internal resistance of shell earth referenced at the point of the connection of ground. For this, the resistor/resistance $Z_{аб}$ must not be turned into zero at frequencies of range, i.e., on must be the resonance of the shell of feeder in the range of the frequencies of the radio direction finder.

I pass to the examination of selection l_n and L_n of antenna-feeder system. It is analogous with that, as we this are made during the analysis of H-shaped system, was decomposed by emf induced in antennas, to phase and nonphase components, we will obtain two equivalent diagrams (Fig. 7.18) for phase and nonphase components of emf. In figure C_n - the capacitance/capacity of field coil on screen, Z_a - input antenna resistance.

Phase currents (single-cycle) are closed through C_n to the earth; nonphase currents (push-pull) are passed through $L_{n\phi}$ (C_n it does not affect).

For the calculation of currents, it suffices to examine the halves of system; thus, diagrams can be simplified (Fig. 7.19).

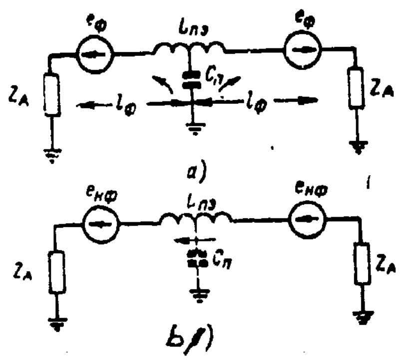

Fig. 7.18. Equivalent diagram of the U-shaped system: a) for the phase component of current; b) for the nonphase component of current.

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On Fig. 7.19a, the resistor/resistance of the capacitance/capacity of order $C_{\text{п}} = 20$ пф is much more than the resistor/resistance of field coil $X_{\text{пэ}}$. Therefore it is possible to consider feeder loaded to capacitance/capacity $\frac{C_{\text{п}}}{2}$. As we this are made in the calculation of H-shaped system, instead of, including $\frac{C_{\text{п}}}{2}$, it is possible to lengthen feeder to value Δl_{Φ}, moreover

$$\rho_{\Phi} \operatorname{ctg} m (l_{\Phi} + \Delta l_{\Phi}) = \frac{\rho_{\Phi} \operatorname{ctg} m_{\Phi} l_{\Phi} - \rho_{\Phi}^2 \omega \frac{C_{\text{п}}}{2}}{\rho_{\Phi} \frac{C_{\text{п}}}{2} \omega \operatorname{ctg} m_{\Phi} l_{\Phi} + 1}.$$

Let us designate $l_{\Phi} + \Delta l_{\Phi} = l'_{\Phi}$.

The condition of a sharp increase in the single-cycle current (from phase component emf) will be equality zero of the reactance of diagram Fig. 7.19a, in reference to points 1-2, i.e.,

$$X_{a\Phi} - \rho_\Phi \operatorname{ctg} m_\Phi l'_\Phi = 0$$

or $$\rho_B \operatorname{ctg} m l_B + \rho_\Phi \operatorname{ctg} m_\Phi l'_\Phi = 0, \qquad (7.64)$$

moreover l'_Φ differs by 3-4o/o from length l_Φ.

If we compare equality (7.64) with the equality

$$\rho_B \operatorname{ctg} m l_B + \rho_\Phi \operatorname{ctg} m_\Phi l_\Phi = 0, \qquad (7.64')$$

that not difficult to conclude that, since $l'_\Phi \approx (1{,}03-1{,}04) l_\Phi$, the frequency, which corresponds to equality (7.64), is lower than frequency which corresponds to equality (7.64'), to 3-4o/o.

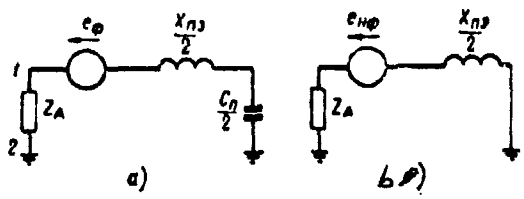

Fig. 7.19. Diagrams of the halves of the U-shaped system: a - for the phase component of current; b - for the nonphase component of current.

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Let us turn to expression for $X_{n\phi}$ (7.45) and (7.46). It is obvious that equality (7.64') coincides on condition $X_{n\phi}=\infty$. Approximately with this same condition coincides the derived by us for circuit Fig. 7.19a resonance condition.

So that the resonance does not lie/rest at working frequency band, it is necessary to satisfy the inequality

$$l_n < \frac{\lambda}{2\pi} \operatorname{arcctg}\left[-\frac{\rho_\phi}{\rho_n} \operatorname{ctg} m_\phi l_\phi\right], \qquad (7.65)$$

moreover λ and m_ϕ they must be undertaken for the frequency, lying to 15-20o/o higher than upper boundary of working frequency band.

Resonance condition for a push-pull current (from the nonphase components of emf of antennas) in diagram Fig. 7.19b coincides on condition of the absence of the resonance of an entire antenna feeder system of the pair of antennas on the basis of which is conducted the selection of inductance L_n. Thus, no new limitations from Fig. 7.19b for the selection of the elements appear.

During the calculation of effectiveness ρ_0 we are guided that which was presented into §7.5.

7.8. Calculation of transformer and balanced systems.

Calculation of transformer system.

Fundamental antenna circuits of system for the cases of applying the asymmetric and symmetrical vibrators are depicted on Fig. 7.20a and b.

We assume that the effective resistance of vibrators can be disregarded.

Let us designate:

$X_a = X_{вх}$ - the input reactance of vibrator;

E_{a_1} and E_{a_2} - emf, induced in vibrators (left and with right);

h_e - the effective height of vibrator;

L_a - antenna inductance of the coil of transformer;

$L_ф$ - the inductance of the feeder coil of transformer;

$$M_т = K_т \sqrt{L_a L_ф}.$$

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The equivalent diagram of system is shown to Fig. 7.20c. Let us convert voltages and the loads of both antennas to points ab and cd of feeders (Fig. 7.20d):

$$\left.\begin{aligned} X_{вх\,э} &= \omega L_ф - \frac{M_т^2 \omega^2}{X_{вх} + \omega L_a} = \frac{\omega L_ф [X_{вх} + \omega L_a (1 - K_т^2)]}{X_{вх} + \omega L_a}, \\ \dot{E}_1 &= j \dot{E}_{a_1} \frac{M_т \omega}{X_{вх} + \omega L_a} = j \frac{M_т \omega h_e E}{X_{вх} + \omega L_a} e^{-jmb}, \\ \dot{E}_2 &= j \dot{E}_{a_2} \frac{M_т \omega}{X_{вх} + \omega L_a} = j \frac{M_т \omega h_e E}{X_{вх} + \omega L_a} e^{jmb}, \end{aligned}\right\} \quad (7.66)$$

where E_1 and E_2 - the voltages, converted in secondary circuit of

transformer from the left and right of vibrators respectively.

Thus, the calculation of transformer system is reduced on the calculation of the H-shaped system which has for each antenna (see Fig. 7.201) entry impedance is designed from the formula

$$X_{A}=X_{\text{вх}\,\phi}=\frac{\omega L_{\phi}\left[X_{\text{вх}}+\omega L_{a}(1-K_{\text{т}}^{2})\right]}{X_{\text{вх}}+\omega L_{\phi}} \qquad (7.67)$$

and effective height is equal to

$$h_{c\text{э}}=h_{\partial}\frac{M_{\text{т}}\omega}{X_{\text{вх}}+\omega L_{a}}, \qquad (7.68)$$

where

$$X_{\text{вх}}=-\rho_{\text{в}}\operatorname{ctg} ml_{\text{в}}.$$

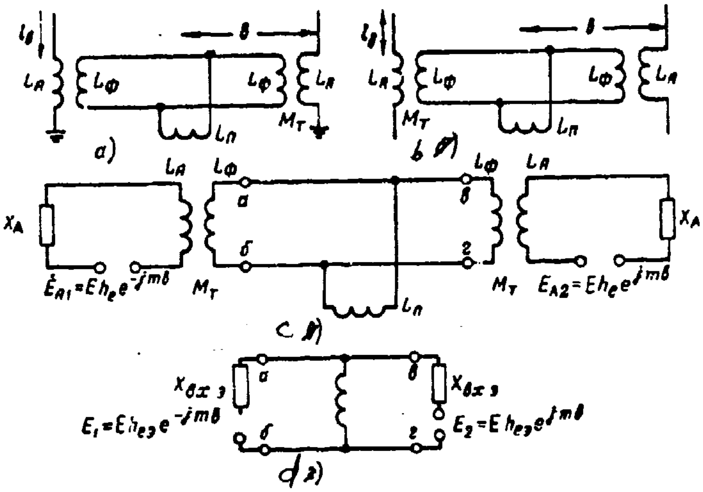

Fig. 7.20. Transformer system of radio direction finder.

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Balanced H-shaped system with the feeders, arrange/located directly on the earth/ground (Fig. 7.21a).

This version of the balanced H-shaped system differs from system examined earlier symmetrical H-shaped in the facts that vertical component of electrical field induces emf only in the upper halfdipoles of vibrators. Therefore during calculation one should consider a series of additional considerations.

The effective height of the single vibrator

$$h_c = \frac{\operatorname{tg}\frac{ml_a}{2}}{m}.$$

The effective height of antenna feeder system from the pair of antennas is computed according to formulas (7.47) or (7.53) with coefficient of 2 in denominator. It is assumed that Z'_3 - the resistor/resistance, which replaces lower halfdipole, is fitted so that on all working frequency band

$$Z'_3 = 0{,}5Z_{nx},$$

where Z_{nx} is complete entry impedance of vibrator, in reference to the points of the switching on of feeders.

The resistor/resistance of vibrator, in reference to the points of the switching on of field coil (Z_{a6}), is computed as H-shaped system, by formulas (7.48), (7.50) and (7.54). As before

$$Z_{nф} = 0{,}5Z_{a6}.$$

In other respects the calculation does not differ from the calculation of symmetrical H-shaped system.

Balanced H-shaped system with feeders, elevated above the earth/ground (Fig. 7.21b).

Emf is induced in upper halfdipoles by length $l_в$ and in lower halfdipoles by length $l_н$. Equivalent (replacing) resistor/resistance $Z_э$ is selected equal to resistor/resistance earth referenced vertical condutor by length $l_в-l_н$ (BV in figure).

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Thus, impedances of halfdipoles of point A are identical. The effective height of the vibrator

$$h_д=h_{дI}+h_{дII}, \qquad (7.69)$$

where $h_{дI}$ - the effective height of upper halfdipole, is determined from the formula

$$h_{дI}=\frac{\operatorname{tg}\frac{ml_в}{2}}{m};$$

$h_{дII}$ - the effective height of lower halfdipole. Taking into account that it is loaded on top and therefore it has approximately uniform current distribution $h_{дII}\approx l_н$.

Then for $h_д$ on the basis (7.69) we obtain with small $ml_в$

$$h_д=\frac{\operatorname{tg}\frac{ml_в}{2}}{m}+l_н.$$

In other respects one should use the method of the calculation of symmetrical H-shaped system.

7.9. Matching devices.

Let us examine devices and the diagrams, which ensure impedance matching of antenna with the resistor/resistance of feeder. The character of matching devices to a considerable degree depends on the width of the section of the frequency band, in which it is required to carry out agreement.

For antenna matching with feeder at the fixed/recorded frequency, mainly in VHF range, are applied single-stub, double-stub and three-stub agreements.

single-stub agreement, or V. V. Tatarinova's jet loop, are applied in the open air feeders.

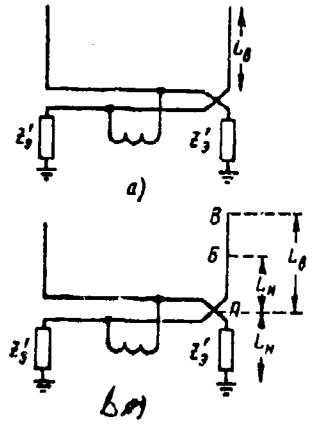

Fig. 7.21. Balanced H-shaped system.

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Short-circuited stub is connected to the feeder of antenna in that place where conductance is equal to $\frac{1}{\rho_\Phi}$ (ρ_Φ- the wave impedance of feeder). The length of loop is selected so as at the points of its switching on to compensate for reactive conductivity of the feeder of antenna. Then feeder proves to be loaded on ρ_Φ.

Double-stub agreement they apply, when the feeder of antenna is carried out in the form of coaxial cable and the movement by it of jet/reactive loop is impossible. Two Pieces of cable (loops) with the

movable short-circuiting pistons are included previously in the cable of feeder at certain distance from each other, for example in $\lambda/4$ or $3/8\ \lambda$, moreover the first piece is included in parallel to the connection of antenna. By a change in the length of the first cable (loop), they attain that the conductance in the place of the connection of the second would be equal to $\frac{1}{\rho_\Phi}$. After this by a change in the length of the second loop is obtained the compensation for susceptance of the feeder of antenna. Double-stub agreement is possible thus far $R_a > \rho_\Phi$ with the distance between loops $\lambda/4$ and $R_a > \frac{1}{2}\rho_\Phi$ with distance $3/8\ \lambda$. With the nonfulfillment of these conditions, additionally is connected at certain distance from the second still the third piece of cable (third loop). The designation/purpose of the first two loops is obtaining in the place of the connection of the third - conductance $\frac{1}{\rho_\Phi}$. Then by the adjustment of the length of the third loop they attain the compensation for the reactance of the load of feeder and agreement of feeder. Three-stub agreement is universal, used during any antenna resistances.

If antenna resistance is at the fixed/recorded frequency effective resistance R_a, then for its transformation into the wave impedance of feeder ρ_Φ can be used the quarter-wave section of feeder with wave impedance $\rho_T = \sqrt{R_a \rho_\Phi}$.

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With the aid of this quarter-wave cut of line, it is possible to obtain the agreement of feeder, also, during any impedance of antenna. For this, the cut of line is connected in the nearest node or voltage antinode. In the first case the resistor/resistance of feeder $R'_{\Phi}=\rho_{\Phi}K_{6B}$ (reactance is equal to zero) and $\rho_{T}=\rho_{\Phi}\sqrt{K_{6B}}$. In the second case $R''_{\Phi}=\frac{\rho_{\Phi}}{K_{6B}}$ and $\rho_{T}=\frac{\rho_{\Phi}}{\sqrt{K_{6B}}}$.

Mono to fit such wave impedance ρ_{T} and the length l_{T} of cut cable, so that it will match at the assigned frequency any resistor/resistance $Z_{H}=R_{H}+jX_{H}$ with any other resistor/resistance $Z_{BX}=R_{BX}+jX_{BX}$. In this case, [8.13]

$$\rho_{T}=\sqrt{A+\frac{BD}{C}}\ \text{and}\ \operatorname{tg} m_{T}l_{T}=\frac{C\rho_{T}}{B},$$

where

$$A=R_{H}R_{BX}-X_{H}X_{BX};$$
$$B=R_{H}X_{BX}+R_{BX}X_{H};$$
$$C=R_{BX}-R_{H};$$
$$D=X_{BX}-X_{H};$$

m_{T} is a phase constant of cable.

The described methods provide the agreement of feeder in narrow

frequency band.

When antenna has constant effective resistance R_a, for its agreement with the wave impedance of feeler in the range of frequencies is applied the cut of line with the smoothly or gradually changing linear parameters, for example exponential feeder or transformer with the concentrated cell/elements. The length of the cut of line with the changing parameters determines the minimum frequency by which the reflection coefficient at the end/lead of the feeder, which goes to receiver, does not exceed the determined permissible value. The transformer it works the, the nearer it to ideal.

Ideal is the transformer which loss-freeta and leakage flux, inductive reactances of its windings the much more transformed resistor/resistances.

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Approach/approximation to these requirements to more easily achieve during the application/use of ferrite core transformers [4.5]. As matching transformer can be used the four-pole from reactive cell/elements. Calculation of this type transformers is given in [7.9].

To expand the frequency band of the agreement of affective resistance with the aid of the quarter-wave section of line is possible by the supplementary switching on of compensating short-circuited stub.

In textbooks on the calculation of antennas, are given more detailed descriptions and the calculations of indicated matching devices [3.2, 3.3, 3.5].

To the questions of the range agreement of impedance from the concentrated cell/elements with effective resistance are ledicated [3.2, 3.3]. In [7.10], is examined the possibility of obtaining in this case with the aid of passive linear four-poles and transformer in the required frequency band of the maximum KBV.

Passive networks serve for the compensation for reactance, and transformer converts the effective resistance of antenna into that which is required. Investigations will show that during the assigned change in the complex resistance maximally attainable KBV and the band of the frequencies of the agreement $f_{макс} - f_{мин} = \Delta f_c$ are mutually connected. Improvement of KBV is led to decrease Δf_c and vice versa. If antenna resistance in certain frequency band the equivalently to consecutive or parallel circuit from cell/elements $L_п$, $C_п$, $R_п$, moreover resonance frequency of duct $f_0 = \sqrt{f_{макс} f_{мин}}$, then matching

condition with the infinite number of cell/elements in matching device they are determined by the equality

$$\frac{\Delta f_c}{f_0} Q = \frac{\pi}{\ln \frac{1 + K_{6в}}{1 - K_{6в}}}, \qquad (7.70)$$

where Q is an energy factor of the duct of the substitution of antenna resistance on frequency f_0.

The higher the antenna field gain, that is less $\frac{\Delta f_c}{f_0}$ with that which was assigned KBV and vice versa.

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Better agreement is obtained if the matching four-pole is catenary with terminal transformer. Examples of the schematics of matching devices are given to Fig. 7.22. The transformation ratio of transformer must be fitted so that at the medium frequency f_0 KBV corresponds to permissible.

The quality of agreement, i.e., KBV and Δf_c at the assigned value of Q, depends on the number of cell/elements in matching device. On Fig. 7.23, is given the dependence attainable KBV on the given bandwidth of the agreement $Q\frac{\Delta f_c}{f_0}$ with the number of cell/elements in the matching four-pole k = 1, 2, 3, and ∞.

As can be seen from Fig. 7.23, the application/use of more than two cell/elements of agreement is led to very insignificant improvement of matching conditions and does not justify itself. In by direction-finding antenna for system it is necessary to apply several identical matching devices, in the number of utilized antennas. In order to facilitate fulfilling the requirement for the identity of the characteristics of all matching devices, it is expedient to manufacture them only from one cell/element L, C (k = 1), allow/assuming certain deterioration in the matching conditions.

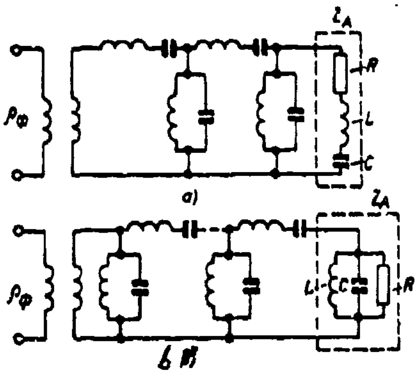

Fig. 7.22. The schematics of the matching devices: a) antenna is equivalent to series connection R, L, C; b) antenna is equivalent to parallel connection R, L, C.

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Are developed the methods of the calculation of the parameters of the ladder network of agreement. For a single-element circuit parameters L and C are designed from the formulas

$$C = \frac{BQ}{R_n 2\pi f_0} \quad \text{and} \quad L = \frac{1}{4\pi^2 f_0^2 C},$$

where B is the coefficient, determined on curve/graph Fig. 7.24.

The device of agreement consists of reactive cell/elements, and therefore it does not lead to power losses in antenna and to deterioration in the relation of the voltages of signal and noise (sensitivity). However, for obtaining in the large band of the frequencies Δf_r of sufficient KBV, it is necessary to available in antenna the small quality Q. The diagram of agreement contains transformer that transforms the effective resistance of antenna circuit. To prepare broadband transformer with large transformation ratio is virtually difficult. Finally, agreement is possible only in that limited frequency band, in which antenna circuit can be replaced with the tuned circuit from the concentrated cell/elements.

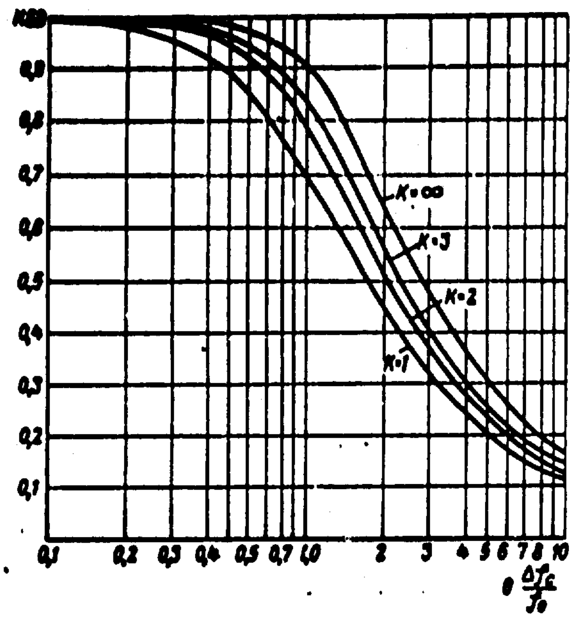

Fig. 7.23. Depenlance of KBV on the given width of the band of agreement with the different number of cell/elements (k).

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So that the resonance frequency of duct would coincide with average frequency range of agreement, can be required the inclusion into the antenna of supplaeentary reactive cell/elements.

For agreement in very wide frequency band for obtaining the low quality Q, can be required the inclusion of supplementary effective resistance into antenna circuit and then the sensitivity of

reception/procedure deteriorates. Deterioration in the sensitivity with the expansion of the band of agreement approximately proportional to square root of the ratio of the expanded band of matching to the band, provided with antenna field gain (see Fig. 7.23).

When is admissible a decrease efficiency in the antenna feeder system, it is possible to also use the diagrams of the agreements, presented in Fig. 7.25 in which according to the idea of their construction is provided the switching on of supplementary effective resistance.

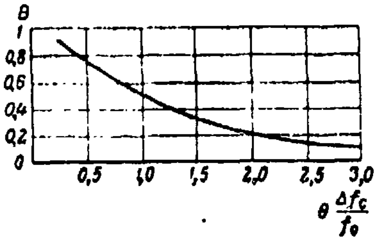

Fig 7.24.

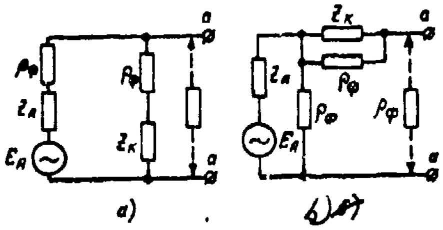

Fig. 7.25.

Fig. 7.24. Dependence of the calculated coefficient B on the given width of the band of agreement.

Fig. 7.25. Diagrams of agreement with the switching on of active cell/elements.

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In these diagrams $Z_к=\frac{\rho_ф^2}{Z_н}$. At the output of the diagrams of the agreement where is connected feeder, resistor/resistance will be:

for circuit a $$Z_{aa}=\frac{(\rho_ф+Z_н)\left(\rho_ф+\frac{\rho_ф^2}{Z_н}\right)}{2\rho_ф+Z_н+\frac{\rho_ф^2}{Z_н}}=\rho_ф.$$

for circuit b $Z_{aa}=\frac{Z_a\rho_\phi}{Z_a+\rho_\phi}+\frac{\frac{\rho_\phi^2}{Z_a}\rho_\phi}{\frac{\rho_\phi^2}{Z_a}+\rho_\phi}=\rho_\phi.$

Assuming that the feeder from receiver is loaded for wave impedance ρ_ϕ, we will obtain for a transmission factor from antenna to the input of the cable

$$k_K=\frac{1}{2}\frac{\rho_\phi}{\rho_\phi+Z_a},$$

i.e. transmission factor decreases two times in comparison with k_K in diagram without agreement. How many times deteriorates the sensitivity of reception/procedure. The indicated diagram of agreement can work in how conveniently wide frequency band.

In [7.12], is described the method of the selection of the cell/elements of the diagram when the loss of sensitivity can be decreased under the condition of admissibility for KBV of the value smaller than unity.

Resistor/resistance $Z_K=\frac{\rho_\phi^2}{Z_a}$ is approximated by the chain/network of the tuned circuits. For the practical target/purposes of sufficiently taking the number of ducts, equal to the rounded relation of maximum operating frequency to the first resonance frequency of the antenna.

If the antenna operates on the section of the frequencies where the active part of the resistor/resistance R_a is changed

insignificantly (usually between two antiresonance), then in parallel chain/network it is possible to include/connect instead of ρ_Φ resistor/resistance R_a and to take $Z_K = \frac{\rho_\Phi^2}{jX_a}$.

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By the switching on of transformer with transformation ratio $n = \sqrt{\frac{\rho_\Phi}{R_a}}$ is achieved agreement. The transmission gain will be

$$k'_K = \frac{1}{2} \frac{R_a n}{Z_a}.$$

From relation $\frac{k'_K}{k_K} = \frac{(\rho_\Phi + Z_a) R_a n}{Z_a \rho_\Phi} = \frac{\rho_\Phi + Z_a}{n Z_a}$ it is evident that the diagram with transformer gives larger transmission factor, than diagram without transformer, if $n > 1$ and $Z_a < \frac{\rho_\Phi}{n-1}$. Diagram with the transformer virtually can work in the range of frequencies with overlap 2-2.5.

Sometimes for agreement is applied the cathode follower to grid of which is connected the antenna, and to the resistive load of cathode - a feeder (§3.14).

7.10. Calculation of the input circuit of receiver for the matched antenna of system.

The equivalent diagram of input circuit is depicted on Fig. 7.26. For the matched antenna output resistance of feeder, connected to the input of receiver, is active and it is equal to the wave impedance of feeder.

For a figure the condition of optimum coupling (will be

$$\omega M_{опт} = z_1 \sqrt{\frac{R_1}{\rho_ф + R_1}} \text{ when } K_{опт} = \sqrt{\frac{1+a^2}{aQ_1}}, \quad (7.71)$$

where $a = \frac{\omega L_1}{\rho_ф}$, $z_1 \approx \sqrt{\rho_ф^2 + \omega^2 L_1^2}$, $R_1 \ll \rho_ф$.

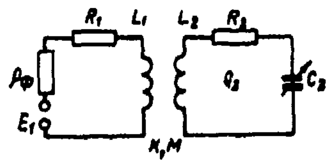

Fig. 7.26. Diagram of input circuit.

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From the viewpoint of obtaining the maximum relation of the voltages of signal and noise (without taking into account of the noise voltage of the first tube) is required larger coupling coefficient (see §7.2). The maximum transmission gain, which corresponds $\omega M_{опт}$, is designed from the formula

$$k_{\text{н макс}} = \frac{1}{2} \cdot \frac{\omega L_2}{\sqrt{(\rho_\phi + R_1) R_2}} \approx \frac{1}{2} \sqrt{Q_2} \sqrt{\frac{\omega L_2}{\rho_\phi}}. \quad (7.72)$$

For an increase $k_{\text{н макс}}$ one should take quality Q_2 possibly larger. The selection of inductance L_2 is determined by frequency band and by adjustable capacitor. Let us determine the permissible value Q_2.

In the tuned circuit from antenna feeder circuit, are introduced

the resistor/resistances

$$\Delta R = \frac{\omega^2 M^2}{\rho_\Phi^2 + \omega^2 L_1^2} \rho_\Phi = \frac{K^2 a}{1 + a^2} \omega L_2, \qquad (7.73)$$

$$\Delta X = -\frac{\omega^2 M^2}{\rho_\Phi^2 + \omega^2 L_1^2} \omega L_1 = -\frac{K^2 a^2}{1 + a^2} \omega L_2. \qquad (7.74)$$

Staggering from action ΔX is determined by the expression

$$\frac{\Delta\omega}{\omega} = \left|\frac{\Delta X}{\omega L_2}\right| = \frac{K^2 a^2}{1 + a^2}$$

or

$$\beta = Q_2 \frac{\Delta\omega}{\omega} = \frac{K^2 Q_2 a^2}{1 + a^2}. \qquad (7.75)$$

In optimum coupling $\beta = \alpha$. (7.75')

Input circuit is disturb/detuned not only due to the effect of the reaction of antenna feeder circuit, but also due to an inaccuracy in the coupling of the block/module/units of variable capacity. The first reason affects much more powerfully. Let us supply requirement that at midband frequency ω_{cp} detuning due to the effect of reaction would be not more the half of the passband of the tuned circuit, i.e., so that would be made the requirement β 0.5 or according to formula (7.75') for a midband frequency

$$a \lessapprox 0{,}5. \qquad (7.76)$$

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Expression (7.76) serves for the selection of the inluctance of the coil of the antenna feeder of circuit:

$$L_1 < \frac{0,5\rho_\phi}{\omega_{cp}}.$$

Entry impedance of circuit from antenna feeder system can be designed by the formula

$$Z_{\text{вх}} = j\omega L_1 + \frac{\omega^2 M^2}{R_2 + \Delta R + j\Delta X} \approx j\omega L_1 + \frac{K^2 Q_2 \omega L_1}{1 + j\beta}$$

or

$$Z_{\text{вх}} = \frac{K^2 Q_2 \omega L_1}{1 + \beta^2} + j\omega L_1 \left(1 - \frac{K^2 Q_2}{1 + \beta^2}\right) = R_{\text{вх}} + jX_{\text{вх}}.$$

In the optimum coupling when are satisfied conditions (7.71) and (7.75'), $R_{\text{вх}} = \rho_\phi$ and $X_{\text{вх}} = 0$.

If due to the imprecise coupling of the block/module/units of variable capacity the inclined grid circuit of the first tube obtains the supplementary detuning, which is characterized by value

$$\Delta\beta = \frac{1}{2} Q_2 \frac{\Delta C}{C}, \qquad (7.77)$$

that this will lead to a change in entry impedance and to the appearance of reflection from input circuit. A change in entry impedance will be

$$\frac{\Delta R_{\text{вх}}}{\rho_\phi}=\frac{\partial R_{\text{вх}}}{\partial\beta}\Delta\beta\frac{1}{\rho_\phi}=-\frac{K^2Q_2}{1+\beta^2}\,\frac{2\beta}{1+a^2}\Delta\beta,$$
$$\frac{\Delta X_{\text{вх}}}{\rho_\phi}=\frac{\partial X_{\text{вх}}}{\partial\beta}\Delta\beta\frac{1}{\rho_\phi}=\frac{1-\beta^2}{1+\beta^2}\,\frac{aK^2Q_2}{1+\beta^2}\Delta\beta.$$

In the optimum coupling

$$\frac{\Delta R_{\text{вх}}}{\rho_\phi}=-\frac{2a}{1+a^2}\Delta\beta,$$
$$\frac{\Delta X_{\text{вх}}}{\rho_\phi}=\frac{1-a^2}{1+a^2}\Delta\beta.$$

Coefficient of reflection of voltage from input circuit in the optimum coupling

$$p=\frac{1}{2}\sqrt{\left(\frac{\Delta R_{\text{вх}}}{\rho_\phi}\right)^2+\left(\frac{\Delta X_{\text{вх}}}{\rho_\phi}\right)^2}=\frac{1}{2}\Delta\beta.$$

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In §4.12 shown, that for the limitation of errors due to the dissimilarity of the electrical lengths of cables the traveling-wave ratio of the load of feeder must be

$$K_{\text{бв}}\geqslant 0{,}4,$$

whence the permissible reflection coefficient was equal to

$$p_{\text{доп}}=\frac{1-K_{\text{бв}}}{1+K_{\text{бв}}}=\frac{1-0{,}4}{1+0{,}4}=0{,}43.$$

The permissible staggering is determined by inequality $\Delta\beta\leqslant 2p_{\text{доп}}$

or $\Delta\beta \leqslant 0.86$.

From (7.77) it follows that energy factor of the tuned circuit must be selected, on the basis of the condition

$$Q_2 < \frac{2\Delta\beta}{\frac{\Delta C}{C}} \text{ or } Q_2 < \frac{1{,}76}{\frac{\Delta C}{C}}.$$

For example, if $\Delta C/C = 0.05$, then $Q_2 \leqslant 1.76/0.05$ and $Q_2 \leqslant 35$.

Thus, we will determine the parameters of all cell/elements, entering input circuit of receiver. Further calculation is conducted by formula (7.72).

7.11. Compensation for antenna effects.

In §§ ~~4.3~~ 4.2, 4.5, 5.3, 6.5 was shown, that on a whole series of reasons in the directed system appear composing emf, out of phase to 90° from the fundamental of emf.

These composing it is possible independent of the reasons, which created them, to compensate for one and the same methods.

The process of the compensation for those composing emf, that differ in phase to 90° from the fundamental of emf, is called the compensation for antenna effects.

With visual direction-finding methods it is not accepted to compensate for antenna effects, since extra-phase emf are not created in them such difficulties as in auditory radio direction finders, and, furthermore, the process of compensation increases time of direction finding.

Antenna effects also less appear in systems with the large separation of antennas, than in cosinusoidal.

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Therefore further the compensation for antenna effects is examined in connection with radio direction finders with the reading of bearing for audition on the minimum and with antenna system with cosinusoidal directional characteristic.

The compensation for antenna effect lies in the fact that in input circuit of the directed system artificially is introduced

supplementary emf, equal in magnitude and opposite on the sign of emf of antenna effect. This supplementary emf can be obtained from the auxiliary antenna or the auxiliary framework. The compensation for antenna effect compulsorily must be that which is controlled, since the value of antenna effect depends in the general case on the wavelength, direction of propagation, time of days and year, and also on other reasons (humidity and the connected with it degree of insulation of antenna and surrounding object/subjects, the upper-air conditions and so forth). To ensure the constant equality of the value of introduced emf and emf of antenna effect during changes in these reasons is virtually impossible.

Regardless of the fact, is applied or is not applied compensation, must be accepted all measures for the elimination of antenna effect, since with its high value are unavoidable supplementary instrument errors.

Introduced for compensation emf must differ in phase from the fundamental of emf to angle, closest possible to 90°. The nonfulfillment of this condition can lead to an increase in the resulting inphase component of antenna effect and, therefore, to an increase in the angle of shift of bearing. For this, explanation let us examine vector diagram (Fig. 7.27).

On diagram E_p designates emf of the framework, E_a is emf of antenna effect, out of phase by the relatively first on angle φ_a.

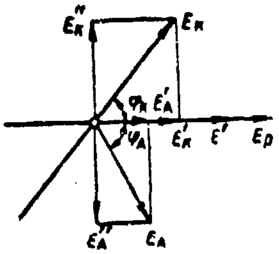

Fig. 7.27. Vector diagram of the compensation for antenna effects.

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It is decomposed $E_{\text{а}}$ to two composing; $E'_{\text{а}}=E_{\text{а}}\cos\varphi_{\text{а}}$ in phase with $E_{\text{р}}$ and $E''_{\text{а}}=E_{\text{а}}\sin\varphi_{\text{а}}$, that out of phase to 90°.

Let phase-compensating emf be assigned and equal to $\varphi_{\text{к}}$. Regulating the value of emf of compensation, we must fit it so that it component/term $E''_{\text{к}}$, shifted relative to $E_{\text{р}}$ on phase to 90°, will be equal and was opposite $E''_{\text{а}}$. In this case $E''_{\text{а}}$ and $E''_{\text{к}}$ they average out and we will obtain clear, clear zero positions. Consequently,

$$E''_{\text{к}}=-E''_{\text{а}}.$$

But

$$E''_{\text{к}}=E_{\text{к}}\sin\varphi_{\text{к}},\quad E'_{\text{к}}=E_{\text{к}}\cos\varphi_{\text{к}}.$$

Hence

$$E_{\text{к}}\sin\varphi_{\text{к}}=-E_{\text{а}}\sin\varphi_{\text{а}},$$
$$E'_{\text{к}}=E_{\text{к}}\cos\varphi_{\text{к}}=-E_{\text{а}}\frac{\sin\varphi_{\text{а}}}{\operatorname{tg}\varphi_{\text{к}}}.$$

Store/adding up now $E'_{\text{к}}$ and E'_{a}, we will obtain the total quantity of cophasal emf, not dependent from the direction of the incident wave:

$$E' = E'_{\text{a}} + E'_{\text{к}} = E'_{\text{a}}\left(1 - \frac{\operatorname{tg}\varphi_{\text{a}}}{\operatorname{tg}\varphi_{\text{к}}}\right).$$

If relation $\frac{\operatorname{tg}\varphi_{\text{a}}}{\operatorname{tg}\varphi_{\text{к}}}$ negatively (as this is represented in figure), then $E' > E'_{\text{a}}$ and the amount of the displacement of the minimum because of compensation will increase. When $\varphi_{\text{к}} = 90^\circ$ $\operatorname{tg}\varphi_{\text{к}} = \infty$, $E' = E'_{\text{a}}$, i.e. the displacement will not increase.

If emf of antenna effect depends on direction (effect of the return emitters), then the condition of the compensation for the nonphase component of antenna effect and the condition of obtaining complete zero with direction finding $E_{\text{р}} \times \sin(p - 0) + E'_{\text{a}} + E'_{\text{к}} = 0$ cease to be independent variables, since amplitude and the phase E_{a} in this case, as it is shown, are also function p. During the different installations of the handle of the compensation for antenna effects, can be found the different positions of the framework, which correspond to the minimum of audibility. Only by method successive corresponding of, revolving simultaneously the handle of the compensation for antenna effects and the framework, it is possible to

find the leepest minimum, which corresponds to bearing.

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With the use of additional antenna for the compensation for nonphase field from return emitter, are possible supplementary errors. Actually, in the process of the compensation for emf from nonphase field, is introduced into the framework the secondary stress, cophasal with emf, induced within the framework by the field of transmitter, which produces error with direction finding. In order to restrict this supplementary error, it is necessary to restrict the permissible value of compensated for nonphase emf.

Let us examine in more detail, to what is led the application/use of a compensative antenna during the elimination of the action of field $E''_{\text{он}}$ (§5.3). The auxiliary antenna affects the field

$$\dot{E}_{\Sigma} = (E + E'_{\text{он}}) + jE''_{\text{он}}. \qquad (7.78)$$

Then, the antenna is strongly detuned and therefore emf within the framework, induced from antenna, coincides in phase with emf, induced in antenna circuit.

Let us designate h_n that which operates high-compensation antenna k_{np} the scaling factor of emf of antenna into voltage within

the framework from antenna. Then the condition of the reading of bearing for zero audibility will be

$$jh_p[E\sin(p-\theta)+(E'_{on}+jE''_{on})\sin(\psi-\theta)]-\\ -k_{ap}h_a[(E+E'_{on})+jE''_{on}]=0. \qquad (7.79)$$

For this fulfillment of equalities, it is necessary that individually the real and imaginary parts of it are equal to zero. Physically equality zero of real part means that the communication/connection of the framework with antenna, determined by coefficient k_{ap}, one should control so as to eliminate nonphase component emf of the framework, induced with the field of return emitter and, thus, to obtain pure/clean of zero audibility. For this, must be

$$k_{ap}h_a=-\frac{E''_{on}\sin(\psi-\theta)}{E+E'_{on}}h_p. \qquad (7.80)$$

After substituting (7.80) in (7.79), we will obtain

$$h_p\left[E\sin(p-\theta)+E'_{on}\sin(\psi-\theta)+\frac{E''^2_{on}\sin(\psi-\theta)}{E+E'_{on}}\right]=0$$

or

$$h_p\left[E\sin(p-\theta)+\left(E'_{on}+\frac{E''^2_{on}}{E+E'_{on}}\right)\sin(\psi-\theta)\right]=0.$$

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Thus, cophasal with the field of transmitter field component of reradiation (E'_{on}) when the compensation for antenna effects is present,

seemingly grow/risen and stops

$$E'_{on_\Sigma} = E'_{on} + \frac{E''^2_{on}}{E + E'_{on}}. \qquad (7.81)$$

The effect of component E'_{on} is developed in the fact that of the radio direction finder, which has the compensation for antenna effects, grow/rises the bearing error. Bearing with the controlled (before obtaining of the complete zero audibility) compensation for antenna effects will differ from bearing without compensation. Limiting maximum supplementary error, i.e., relation $\frac{E''_{on}}{E}$, thereby in accordance with (7.80) we limit value k_{ap}.

If we suppose that $E'_{on}=0$ and to allow during compensation supplementary error Δ = 2°, then of formula (7.81) follows that $\Delta \approx \left(\frac{E''_{on}}{E}\right)^2 \approx 0{,}03$ or $\frac{E''_{on}}{E} = \sqrt{0{,}03} = 0{,}173$. Therefore must be made condition $k_{ap} h_a < 0{,}173 h_p$, which limits maximum permissible emf of the compensation for antenna effects.

The circuit diagram of the compensating open antenna is represented on Fig. 7.28. The communication/connection of antenna with the circuit of the framework is here undertaken inductive. The value of communication/connection is regulated with the aid of variometer L_a, L_1, L_2. Its winding, the circuital of the framework, is divided to two parts L_1 and L_2, connected round trip of the framework for the preservation/retention/maintaining of its symmetry.

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During the practical fulfillment of the compensating antenna, must be observed the also following conditions:

1) antenna must not have direct/straight communication/connection with the framework;

2) antenna must be vertical or weakly inclined. Significant horizontal parts can cause supplementary quadrantal deviation, since antenna with the developed horizontal part can be considered as capacitive duct;

3) antenna must be weakly connected with the framework in order to eliminate the effect of a change in the communication/connection for the tuning of the framework;

4) on short waves in order to avoid the appearance of those who are changing with a change in the wave of phase differences of emf of the framework and antenna (due to the effect of increment), antenna is assembled in the center of frame system.

Besides the examined inductive coupling of antenna with the framework, is possible also the application/use of capacitive coupling.

The diagram of the compensation for antenna effects in goniometric system in no way differs from diagrams in system with rotatable loop, only everything that which was discussed the frame duct, here it is related to the duct of the selector of goniometer.

It should be noted that with respect to a change in the value of the required compensation for antenna effects it is possible approximately to judge the character of the reason, which caused diffuseness of the readings of bearing.

Let us give several examples.

1. In presence only of its own antenna affect and direct reception/procedure, sign of required communication/connection with auxiliary antenna for elimination of diffuseness of bearing is not changed during rotation of framework through 180°. The value of required communication/connection remains almost constant, independent of the direction of the oriented radio station (with the constancy of wavelength).

2. For ship radio direction finder usually is obtained dependence of communication/connection, required for compensation for antenna effects (diffuseness of bearing), with auxiliary antenna on direction of oriented radio station, close to semi-circular. This is explained to the facts that the diffuseness of bearing is caused mainly by reradiation of the masts, tubes and other object/subjects, analogous in action to detuned antenna. to the maximum of communication/connection usually they are observed with the direction finding of the radio stations, arrange/located perpendicularly to the center-line plane of the ship, in which is arrange/located return emitter.

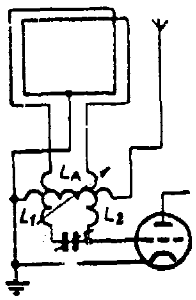

Fig. 7.28. Circuit diagram of compensative antenna.

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3. Value of required communication/connection with auxiliary antenna at presence from ionosphere of waves reflected usually sharply is changed in time.

4. If auxiliary antenna is insufficiently detuned and oriented field elliptical, then compensation for diffuseness can bring to obtaining false minimums with direction finding.

5. During compensation for large antenna effects, can appear bearing errors (7.81).

7.12. Calculation of the cell/elements of the compensation for antenna effects.

In practice most frequently is applied auxiliary detuned antenna, the inductively circuital framework (Fig. 7.28). The equivalent circuit of the communication/connection of the framework with antenna is shown to Fig. 7.29.

On equivalent diagram are designated:

L_1 - inductance of the framework;

L_2 - inductance of the winding of the variometer, connected in duct framework; for simplicity coil L_2 is not divided to two as to Fig. 7.28;

L_3 - inductance of the winding of variometer, connected in antenna;

C_1 - the capacitance/capacity of the duct of the framework;

Z_a - equivalent antenna resistance (on Fig. 7-29 C_2);

C_3 - the capacitance/capacity of antenna lead-in;

Eh_n is emf in antenna.

In calculation we will disregard the effective resistance of antenna circuit, taking into account its detuning.

Therefore $Z_a \approx jX_a = -j\rho_a \operatorname{ctg} ml_a$, where ρ_a, l_a is wave impedance and the geometric length of antenna.

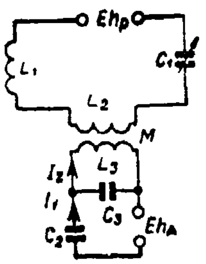

Fig. 7.29. Equivalent diagram of the communication/connection of the framework with antenna.

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In further formulas we proceed from the direct connection of antenna to input circuit of the framework. If antenna is connected to the radio direction finder through the long feeder, then one should convert emf in antenna and its resistor/resistance for the end/lead of the feeder the formulas §7.5.

In the majority of cases, the antenna is taken small length in comparison with the minimum wavelength of direction finding and it is possible to count which X_a is capacitance, i.e.,

$$X_a = -\frac{1}{\omega C_2}.$$

In calculations of auxiliary antenna we proceed from this prerequisite/premise.

Let us designate parallel connection L_3, C_3 by X_2:

$$jX_2=\frac{j\omega L_3\frac{1}{j\omega C_3}}{j\omega L_3+\frac{1}{j\omega C_3}}=j\frac{\omega L_3}{1-\omega^2 L_3 C_3}.$$

Then (Fig. 7.29)

$$I_1=\frac{Eh_a}{(X_2+X_a)},\ I_3=\frac{I_1 X_2}{\omega L_3}=\frac{Eh_a X_2}{\omega L_3 (X_2+X_a)},$$

but

$$\frac{X_2}{X_a+X_2}=\frac{X_a X_2}{X_a+X_2}\frac{1}{X_a}=\frac{\omega L_3}{\beta}\frac{1}{X_a}=-\frac{\omega^2 L_3 C_2}{\beta},$$

where $\beta=1-\left(\frac{\omega}{\omega_a}\right)^2$;

$\omega_a=\frac{1}{\sqrt{L_3(C_2+C_3)}}$ - resonance frequency by the antenna of circuit.

Frequency ω_a usually is taken $\omega_a < 0.7\omega_{мин}$ or $\omega_a > 1.4\omega_{макс}$, where $\omega_{мин}$ and $\omega_{макс}$ - the minimum and maximum frequencies of range.

With the assigned antenna of capacitance/capacity C_2 and C_3, are known. Being assigned ω_a, it is possible to determine

$$L_3=\frac{1}{\omega_a^2 (C_2+C_3)}.$$

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We convert expression for the amplitude of current I_2:

$$I_2 = Eh_a \frac{\omega C_2}{\beta}.$$

Emf, induced with current I_2 in the duct of the framework, will be

$$E_{ap} = \omega M_{ap} I_2 = Eh_a \omega^2 M_{ap} C_2 \frac{1}{\beta}, \qquad (7.82)$$

where $M_{ap} = K\sqrt{L_2 L_3}$;

K is a coupling coefficient between coils L_2 and L_3 variometer; it is determined by the selected construction of variometer.

Let us designate the ratio of emf of antenna effects to emf from the frame reception/procedure by α: then must be made the requirement

$$E_{ap} \leq \alpha E h_p, \qquad (7.83)$$

where for a ship radio direction finder usually α = 0.1-0.2; for the coastal α = of 0.05-0.1.

From formulas (7.82) and (7.83) is determined inductance L_2:

$$L_2 \geqslant \left(\frac{\alpha}{K}\right)^2 \left(\frac{h_p}{h_n}\right)^2 \frac{1}{\omega^4} \frac{1}{C_9^2 L_1} \left[1 - \left(\frac{\omega}{\omega_n}\right)^2\right]^2.$$

Substituting expression for L_3 and taking into account that $\frac{\omega}{\omega_n} = \frac{\lambda_n}{\lambda}$, we will obtain final calculation formula for inductance L_2:

$$L_2 \geqslant \left(\frac{\alpha}{K}\right)^2 \left(\frac{h_p}{h_n}\right)^2 \left(\frac{C_1 + C_2}{C_1}\right)^2 \left[\left(\frac{\lambda}{\lambda_n}\right)^2 - 1\right]^2 L_3. \quad (7.84)$$

Calculation is conducted for the wave of the range of radio direction finder, most differing from its own wave antenna λ_n.

In the case of calculation L_2 and K for a goniometric system in the communication/connection of antenna with the duct of search coil, one should instead of h_p substitute $p\,\delta_n$, where p is effectiveness of goniometric system; δ_n is circuit damping of selector.

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Usually $L_2 \leqslant 0.1L_1$; then the effectiveness of frame reception/procedure falls insignificantly from switching on L_2 consecutively with L_1. If L_2 is obtained large, then one should increase h_n and produce translation, after assigning $L_2 = 0.1L_1$ and value β. From (7.84) is determined that h_n and then is designed altitude of required antenna.

Let us determine the detuning of frame duct due to the effect of antenna. The introduced from by the antenna of circuit into frame duct reactance is determined from the equation

$$\Delta X=\frac{\omega^2 M_{ap}^2}{\omega L_1 \beta'}=\omega\frac{K^2}{\beta'}L_2,\ \beta'=1-\left(\frac{\omega_a}{\omega}\right)^2.$$

A relative change in the reactance of the framework will be determined from the expression

$$\frac{\Delta X}{X}=\frac{K^2}{\beta'}\frac{L_2}{L_1}.$$

The detuning of frame duct in frequency will be

$$\frac{\Delta f}{f}=\frac{1}{2}\frac{\Delta X}{X}=\frac{1}{2}\frac{K^2}{\beta'}\frac{L_2}{L_1}.$$

So that during the compensation for antenna effects does not change the audibility of the oriented radio station due to detuning, value Δf must be less than 0.5 passbands of the duct of the framework. Calculation Δf one should produce for a wave, closest to λ_a. If $\Delta f/f$ it is great, it comes to increase $|\beta'|$ and to produce the translation of the cell/elements of the diagram of the compensation for antenna effects.

7.13. Calculation of unidirectional reception/procedure.

Cosinusoidal radiation pattern makes it possible to determine two values of bearing, which differ by 180°. For determining single-valued bearing, is applied the combined reception/procedure into the framework and the open antenna with cardioid radiation pattern (§3.8).

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For obtaining the diagram of reception/procedure in the form of cardioid, it is necessary to obtain current in antenna by that cophasal with emf in it. This phase coincidence is obtained precise, if antenna circuit is inclined into resonance with incident wave, or approximately, if in circuit of antenna included high effective resistance. Diagram with the tuned antenna due to its complexity will not find use in radio direction finders.

Are examined below diagrams for the single-valued determination of bearing with auditory direction-finding method. The patterns of the determination of single-valued bearing with visual direction-finding methods are examined during the description of visual direction-finding methods.

Diagram with the unadjusted vertical wire antenna and spiral loop.

In diagrams with untuned antenna, it is not possible to obtain accurately the phasing of emf, induced in the circuit of the framework from antenna, and emf, induced within the framework it is direct. Let us examine, which degree of the disagreement of the phases between these of emf is permissible from practical point of view.

In the circuit of the framework, operate two electromotive forces: emf E_p, induced in it it is direct,

$$\dot{E}_p = -jh_p \mathrm{E} \cos\theta$$

and emf E'_a, induced in it by antenna,

$$\dot{E}'_a = j\gamma h_a \mathrm{E} (\cos\varphi \pm j \sin\varphi).$$

In last/latter formula we assume that the means of communication of antenna with the framework causes phase displacement of emf E'_a relative to current in antenna to angle $\pi/2$. Real factor $\gamma = \frac{|E'_a|}{|E_a|}$ represents the ratio of the module/modulus of emf, induced by antenna within the framework, to emf in antenna itself. φ is a phase angle of current in antenna relative to emf in it. The resulting quantity of emf within the framework will be

$$\dot{E}_{pea} = \dot{E}_p + \dot{E}'_a = j\mathrm{E} [- h_p \cos\theta + \gamma h_a \cos\varphi \pm j\gamma h_a \sin\varphi].$$

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The module/modulus of this value

$$E_{\text{рез}} = \mathrm{E}\sqrt{(-h_p \cos\theta + \gamma h_a \cos\varphi)^2 + (\gamma h_a \sin\varphi)^2} =$$
$$= \mathrm{E}h_p\sqrt{(-\cos\theta + \alpha\cos\varphi)^2 + (\alpha\sin\varphi)^2},$$

where $\alpha = \frac{\gamma h_a}{h_p}$ is relation of the amplitudes of emf, induced within the framework through antenna circuit it is direct. The amplitude of resulting emf at $\theta = 0$ and $\theta = 180°$ will be

$$E_{\text{рез}\,0} = \mathrm{E}h_p\sqrt{(\alpha\cos\varphi - 1)^2 + (\alpha\sin\varphi)^2},$$
$$E_{\text{рез}\,180°} = \mathrm{E}h_p\sqrt{(\alpha\cos\varphi + 1)^2 + (\alpha\sin\varphi)^2}.$$

The relation of these two values characterizes the difference in audibility during the rotation of the framework through 180° and, therefore, clearness of the determination of side. This sense he is called the coefficient of the unidirectional reception/procedure

$$\xi = \sqrt{\frac{(\alpha\cos\varphi + 1)^2 + \alpha^2\sin^2\varphi}{(\alpha\cos\varphi - 1)^2 + \alpha^2\sin^2\varphi}} = \sqrt{\frac{1 + \alpha^2 + 2\alpha\cos\varphi}{1 + \alpha^2 - 2\alpha\cos\varphi}}. \quad (7.85)$$

Value ξ (in dB) as a function of ϕ and α is represented in Fig. 7.30.

Usually they accept, that for the clear determination of side the difference of audibility during the rotation of the framework

through 180° must be not less than 10 dB. From curve/graph it is evident that this difference in audibility can be provided with $\alpha > 0.6$. With this than is more α, the greater phase displacement can be allowed. So, for $\alpha=0.6$ $\varphi_{макс}=20°$; for $\alpha=0.8$ $\varphi_{макс}=30°$ and for $\alpha=1.0$ $\varphi_{макс}=35°$. With a further increase α ξ it decreases. Therefore one ought not to select $\alpha > 1$.

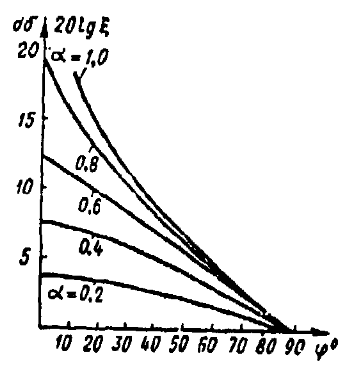

Fig. 7.30. Clearness of determination of side depending on phase displacement.

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The patterns of the determination of side with the use of

untuned antenna are represented in Fig. to 7.31 (7.31a - the diagram of the direct coupling of antenna with the framework, 7.31b - the diagram of inductive coupling).

After designating C_{a} antenna capacity, $C_{\text{в}}$ the capacitance/capacity of antenna lead-in, $R_{\text{п}}$ the resistor/resistance of framework, C the capacitance/capacity of the tuning of the duct of the framework, from the calculation of equivalent diagram 7.32a, we will obtain

$$\gamma = \frac{C_{\text{a}} \sin \varphi}{C_{\text{a}} + C_{\text{в}} + C}, \qquad (7.86)$$

$$a = \frac{C_{\text{a}} \sin \varphi}{C_{\text{a}} + C_{\text{в}} + C} \frac{h_{\text{a}}}{h_{\text{p}}}, \qquad (7.87)$$

$$\operatorname{tg} \varphi = \frac{C_{\text{a}} + C_{\text{в}} + C}{\omega R_{\text{п}} C (C_{\text{в}} + C_{\text{a}})}. \qquad (7.88)$$

Knowing the parameters by the antenna of circuit selecting α and φ, according to that which was presented above it is possible to find $R_{\text{п}}$ and ξ. Calculation is performed on longest wave of the range, where obtaining one-sided reception/procedure presents greatest difficulties.

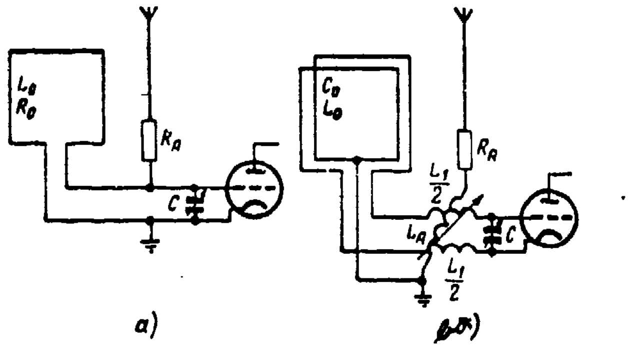

Fig. 7.31. Patterns of determination of side: a) direct switching on of antenna; b) antenna coupling.

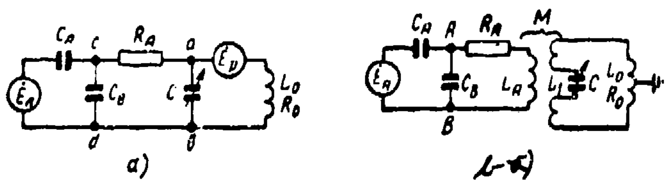

Fig. 7.32. Equivalent patterns of determination of side: a) direct switching on of antenna; b) inductive antenna coupling.

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Equivalent diagram for the case of the inductive coupling of antenna with the framework is given in Fig. 7.32b.

In order to exclude resonance by the antenna of circuit from working wave band, antenna inductance of circuit should select, on the basis of the condition

$$\omega L_a < \frac{1}{\omega (C_a + C_в)}. \qquad (7.89)$$

During satisfaction of this condition, disregarding ωL_a in comparison with $1/\omega(C_a + C_в)$, we obtain the calculation formulas

$$\mathrm{tg}\,\varphi = \frac{1}{\omega R_a (C_a + C_в)}, \qquad (7.90)$$

$$\gamma = -\omega^2 M C_a \sin\varphi, \qquad (7.91)$$

$$\alpha = \frac{h_a}{h_р}\omega^2 M C_a \sin\varphi. \qquad (7.92)$$

Course of computation is following: they determine that L_a, after substituting in (7.89) the greatest frequency of operating range $\omega_{макс}$. Is determined the coefficient of mutual inductance, for which are assigned by the inductance of coupling coil in the circuit of framework L_1 and by the coupling coefficient K. In order not to cause a noticeable decrease in the inductance (and in the effective height) of the framework, they accept L_1 = (0.05-0.1) L_0. Coupling coefficient is selected within limits $K_{макс} = 0.2 \div 0.4$. Calculation they conduct for $K = (0.5-0.7) K_{макс}$. Further, after assigning x, by formula

(7.92) are found sin ϕ.

This determination is conducted for the lowest frequency of range. Knowing ϕ, by formula (7.90) they find that R_n. Finally, through Fig. 7.30 or (7.85) they find ε. If value ε for lowest frequency is satisfactory, is conducted checking ε at the higher frequency of range.

Comparing formulas for the circuits of direct and inductive connection of antenna with the framework, it is possible to see that the direct coupling provides high value γ, which is favorable. On the other hand, the diagram with inductive coupling is more sending. Specifically, in work in several partial ranges in circuit with inductive coupling to more easily ensure optimum conditions in all partial ranges by switching coil L_a.

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In the insufficient value of emf from antenna and the impossibility to increase its height/altitude applies in the channel of antenna amplification. Due to the complications of diagram and control, that appear during the application/use of the tuned circuit in antenna circuit, they are limited to the application/use of a cascade/stage of aperiodic amplification.

Diagram with the unadjusted vertical wire antenna and the unadjusted framework.

If the framework is not inclined, then emf store/add up in the tuned circuit in which operate emf, induced from the circuit of the framework and from antenna circuit. Figures 7.33 gives the diagram in which the antenna is not also inclined and it is inductively circuital. It is obvious that in this case it is possible in equal the case it is possible in equal measure to use any of the systems described above of antenna coupling.

Emf in duct, induced from the circuit of the framework E'_{n} in this case will be out of phase relative to field to angle $90^{\circ}-\phi_{0}$, since is obtained supplementary phase displacement ϕ_{0} in the circuit of the unadjusted framework. Phase displacement ϕ between two emf, that operate in duct, will be equal to a phase difference ϕ_{0} and φ_{a}. where φ_{a} determines the phase of emf, induced in duct from antenna. Thus, in formulas (7.86)-(7.88) and (7.90)-(7.92) hearth ϕ should understand resulting phase displacement:

$$\varphi = \varphi_{a} - \varphi_{0}.$$

The amplitude of emf E'_p in duct will also differ from the amplitude of emf within the framework. Specifically, value E'_p one should substitute in formula (7.85) instead of E_p.

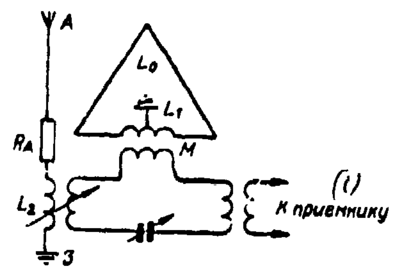

Fig. 7.33. Pattern of determination of side in the case of unadjusted framework.

Key: (1). To receiver.

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For the circuit in question, disregarding the capacitance/capacity of the framework, we will obtain

$$I_p = \frac{E}{R_0 + j\omega(L_0 + L_1)} = \frac{Eh_p \cos\theta}{\omega(L_0 + L_1)} \frac{1}{j + \frac{R_0}{\omega(L_0 + L_1)}} =$$

$$= \frac{Eh_p \sin\varphi'_0}{\omega(L_0 + L_1)} (\cos\varphi'_0 - j\sin\varphi'_0)\cos\theta.$$

where $\operatorname{tg} \varphi'_0 = \frac{\omega (L_0 + L_1)}{R_0}; \varphi'_0 = 90° - \varphi_0.$

Emf in duct will be equal to

$$\dot{E}'_p = \frac{Eh_p \sin \varphi'_0 M}{L_0 + L_1} (\sin \varphi'_0 + j \cos \varphi'_0) \cos \theta.$$

Module/modulus of this value

$$E'_p = \frac{Eh_p M}{L_0 + L_1} \sin \varphi'_0 \cos \theta.$$

Presented in this paragraph directly applicably to the goniometric system which, as shown, is completely equivalent to the framework with inductive coupling.

The addition of emf in separate duct in the majority of cases is applied in goniometric systems with the framework and with the spaced antennas, and also in system with the revolving spaced antennas.

During as the antenna of the system of radio direction finder

one shoull take measures for that so that between electromagnetic field, which operates on the directed system and the omnidirectional antenna, would not be the noticeable phase difference, which was being changed with frequency. This is especially important on ultrashort-wave and skip bands. Therefore, for example, in the ship goniometric frame radio direction finder of short waves auxiliary antenna is assembled in the center of frame system [8.16]. Further, with reference by the antenna of system it is necessary so to select the diagram of input circuits so that the phase of circuital currents of addition from auxiliary antenna and from the directed system changes in the range of frequencies according to identical law.

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For simplification in the process of determining the side, is possible the application/use for the unidirectional reception/procedure of the separate framework, arrange/located perpendicular to fundamental and that which is included in diagram instead of it during transition to the determination of side. In this case, the minimum and the maximum of unidirectional reception/procedure are observed in the same position of the framework, as the minimum with direction finding. The supplementary framework usually is made less than the fundamental, which facilitates the selection of the relationship/ratio of the values of

directed and nondirectional emf at small size/dimensions of the open antenna.

Analogous with this in goniometric system is possible the application/use of supplementary search coil, perpendicular to fundamental and utilized only for determining side.

Use of an antenna effect of the framework and simplified circuits of unidirectional reception/procedure.

The antenna effect of the framework, examined in chapter 5, is equivalent to introduction into its circuit of supplementary emf, which does not depend on direction. With its sufficient value and correct phase relationship, can be obtained cardioid pattern of reception/procedure without supplementary antenna. For this, it is required by the selection of the corresponding diagram to increase antenna effect so so that its emf would be equal to emf within the framework in amplitude and it coincided in phase with the latter. Figures 7.34 gives this diagram.

Midpoint by the antenna of the system through self-inductor L and resistor/resistance R_a is connected with the earth/ground.

Through this coil occur/flow/lasts the full current of antenna effect. Coil L by one of previously described methods is connected with the duct of the selector of the goniometer or framework. This diagram is applied more frequently with the spaced antennas than with the locked framework, since for the first antenna effect is relatively higher.

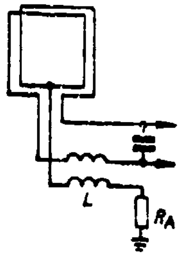

Fig. 7.34. Use of its own antenna effect for determining side.

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Use of the grounded point of the field coils of goniometer in a system with the spaced antennas.

For attendant reception/procedure, the peaking of the minimum with auditory direction finding (compensation for antenna effects) and the determinations of the side of radio stations in radio direction finder instead of the omnidirectional antenna it is possible to utilize a single-cycle current of reception/procedure to the directed system and thus to manage without itself the omnidirectional antenna.

Let us examine the possibility of using the current through the ground wire of the field coils of goniometer for the nondirectional reception/procedure. In the case of the winding/coil of goniometer by star the grounded point is the common point of field coils. In the case even number of antennas, symmetrical winding/coil of field coils and use of each coil for the connection of the opposite pairs of antennas the grounded point must be midpoint of field coils.

Earlier we will establish that for emf of the m antenna it is possible to write (3.50)

$$E_m = E_0 h_{e0}\left[J_0(a) + 2\sum_{p=1}^{\infty} j^p J_p(a)\cos p(\theta - \delta m)\right],$$

where $a = \frac{2\pi}{\lambda} b\cos\beta$; $\delta = \frac{2\pi}{n}$.

During the use of a common point of field coils, all emf of antennas are connected in parallel. Therefore total emf will be

$$\dot{E}_{s} = \sum_{m=1}^{n} E_m = E_0 h_{e0}\left[nJ_0(a) + 2\sum_{p=1}^{\infty} j^p J_p(a) \sum_{m=1}^{n} \cos p(\theta - \delta m)\right]. \tag{7.93}$$

We convert (7.93), taking into account that

$$\sum_{m=1}^{n} \cos\left(\frac{2\pi}{n} pm\right) = n \text{ при}^{(1)} \ p = kn,$$
$$\sum_{m=1}^{n} \cos\left(\frac{2\pi}{n} pm\right) = 0 \text{ для}^{(2)} \text{ других } p,$$
$$\sum_{m=1}^{n} \sin\left(\frac{2\pi}{n} pm\right) = 0 \text{ всегда}^{(3)}.$$

Key: (1). with. (2). for another p. (3). always.

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Then with even number of antennas

$$\dot{E}_{s} = E_0 h_{e0}\left[nJ_0(a) + 2(-1)^{\frac{n}{2}} J_n(a)\cos n\theta + \right.$$
$$\left. + 2(-1)^n J_{2n}(a)\cos 2n\theta + \ldots\right].$$

For six and more antennas usually it is possible to disregard $J_n(a)$, $J_{2n}(a)$ so forth in comparison with $J_0(a)$. Therefore approximately

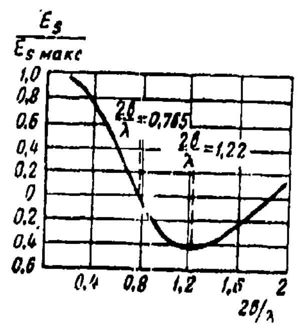
$$E_{s} = E_0 h_{e0} n J_0(a).$$

Maximum value E_s is obtained with a = 0, i.e., with 2b $\rightarrow$ 0. In Fig. 7.35 is constructed the dependence $\frac{E_s}{E_{s\,макс}}$ on 2b/λ with β = 0. From figure we see that when $\frac{2b}{\lambda}=0{,}765$, $E_s=0$ and when $\frac{2b}{\lambda}=1{,}22$ $E_s=-0{,}4E_{s\,макс}$.

Thus, if is placed the requirement to utilize midpoint of field coils for the nondirectional reception/procedure, then it is necessary to satisfy requirement 2b/λ < 0.765.

With the larger separation of antennas, i.e., when 2b/λ > 0.6-0.7, it is possible to select for the nondirectional reception/procedure each time the pair of antennas, which lie/rests at the plane, perpendicular to the direction of the oriented radio station, but diagram and work itself on direction finding in this case strongly become complicated.

Fig. 7.35. Change in emf of undirected action depending on separation of antennas.



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Chapter 8.

VISUAL RADIO DIRECTION FINDERS.

8.1. Fixed loop radio compasses.

Radio direction finders with visible reading of bearing are divided into two groups:

1) direction finders with manual installation by the antenna of system on bearing from readings (§8.1);

2) the direction finders in which the bearing can be counted off on its image or on the position of the arrow/pointer of instrument immediately after tuning to radio station without the need for revolving antenna system (§8.2-8.8).

Are carried out works on the automation of averaging and removal of bearings in the direction finders of the second group. The averaged bearing is developed on digital signal panel (§8.9).

The visual radio direction finders, which determine bearing by the installation of the directional antenna toward the minimum of the depth of modulation, will be called the name fixed loop radio compasses (see Fig. 2.17).

Fixed loop radio compasses widely are applied in aircraft. This is explained by the following reasons:

1) by high interference level on aircraft (including acoustic), which are led to an increase in the subjective error with the taking of bearing by audition in the minimum;

2) by simplicity of the determination of bearing during the use of a visual display in radio direction finder; hence by the possibility of the maintenance of direction finder by the less qualified operators, that especially importantly on aircraft;

3) by simplicity of realization on the fixed loop radio compass of flight to radio station.

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Overall sizes by the antenna of system are limited due to mounting conditions on aircraft; therefore in fixed loop radio compasses is applied as by the antenna of system almost exclusively the combination of the framework and small omnidirectional antenna. For the size decrease by the antenna of system frequently are applied the framework with ferromagnetic core. Voltage of one of the antennas undergoes commutation or modulation and store/adds up with the voltage of the second antenna.

It is expedient switchings to produce in the circuit of the framework, since then in flight to radio station (torque/moment of bearing) is absent the modulating voltage and is feasible the reception/procedure of radio station without interference from modulation. The dissimilarity of diodes or triodes of switching is led during switching of the framework only to different instrument sensitivity during deviation to the right and to the left from course. During switchings of antenna, this dissimilarity causes with direction finding also errors Δ_h. Figures 8.1 gives the resulting diagrams of the reception/procedure when switch is located in antenna circuit and voltages from antenna change their value during switching of phase. The position of bearing is determined with error Δ_h.

The commutation of the framework can be produced electromechanical or with electronic method. Due to the inconstancy

of contacts, short service life and limited switching rate of electromechanical switches standard method is the electronic switching. The devices with the aid of which is realized commutation with sinusoidal envelope shape or modulation with the suppression of carrier frequency, they are called the balanced modulators. In the balanced modulators it is possible to utilize diodes (Fig. 8.2), electron-tube (Fig. 8.3) and transistors, and also multigrid tubes.

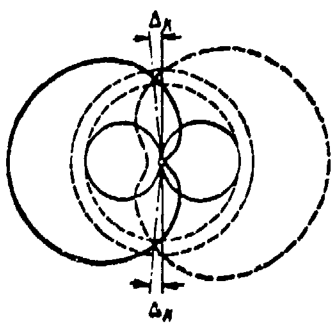

Fig. 8.1. Displacement of bearing during switching of antenna.

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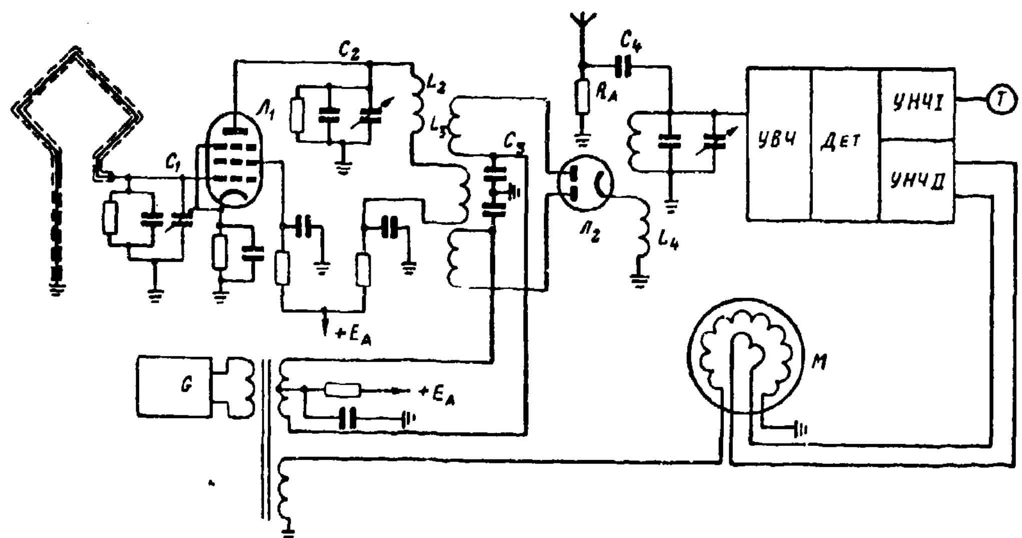

Fig. 8.2. Schematic diagram of fixed loop radio compass with use of diodes.

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For an increase in the sensitivity, taking into account the small effective height of the framework and the increased inherent noise level of tubes of the balanced modulator, between the framework and the balanced modulator usually is placed amplifier stage (Fig. 8.2). In the channel of antenna, it is ossible, as a rule, to manage without itself amplification.

At the output of the balanced modulator, occurs the addition of the modulated voltage of framework and voltage of antenna, which restores carrier frequency. The resulting voltage will be feed/conducted to receiver (usually the superheterodyne type), that contains the cascade/stages of the amplification of the high and intermediate of frequencies, detector and low-frequency amplifier. The output voltage of receiver, which has modulation frequency, supplies the rotor of indicator M, and the voltage of the local oscillator is its stator. Indicator is ferrodynamic instrument.

Readings is determined by currents in stator I_{cr} and rotor I_p and by phase displacement between them ψ:

$$\alpha = k_n I_p I_{cr} \cos\psi. \qquad (8.1)$$

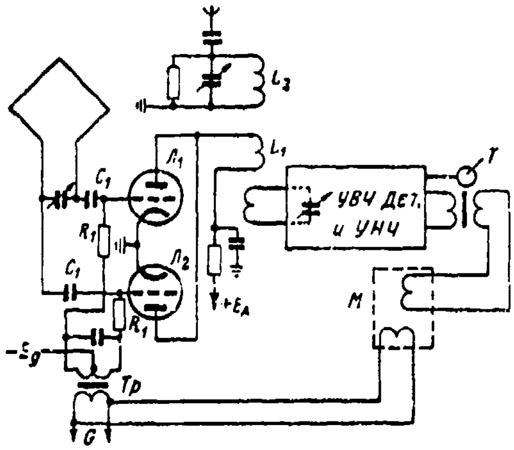

Fig. 8.3. Schematic diagram of fixed loop radio compass with the use of triodes.

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Phase displacement of currents in rotor and stator must be

small, otherwise the throw of pointer decreases and sensitivity falls.

In formula (8.1) it is assumed that currents $I_{\text{р}}$ and $I_{\text{ст}}$ one and the same frequency. If frequency $I_{\text{р}}$ differs from frequency $I_{\text{ст}}$, appears the alternating/variable torque of rotor and rifleman/gunner it comes into oscillatory motion. Because of the inertia of the movable system of frequency indicator, they will decrease with an increase in the frequency, i.e., difference in the current frequencies, which feed rotor and stator. The difference in the frequencies, by which the amplitude of oscillations falls to 0.707 from the maximum, determines the band of the susceptibility of indicator. The band of the susceptibility of indicators composes several hertzes.

Let us examine the work of the balanced modulator in which are used the vaccum tube triodes. The fundamental conclusions which will be obtained, are retained for other circuits of the balanced modulators.

To the grids of the tubes of the balanced modulator $Л_1$ and $Л_2$ will be feed/conducted in antiphase the voltage of the local oscillator of frequency $\Omega/2\pi = F$. The voltage of the framework will be feed/conducted to the grids of tubes also antiphase, and their

anode currents store/add up cophasally. Is possible another construction of the diagram when the grids of tubes $Л_1$ and $Л_2$ are supplied from the framework cophasally, but the anodes are included antiphase, i.e., their currents are deducted. However, the second version is less favorable (see page 422).

Let the current of tubes $Л$ obey the law (work occurs without cutoff)

$$i = I_0 + a_1 e_g + a_2 e_g^2 + a_3 e_g^3 + \ldots,$$

where e_g is a grid voltage; a_1, a_2, a_3 are coefficients of the characteristic of tubes.

Let us designate $E_g \sin \Omega t$ voltage from generator G of low frequency, $E_{р\,макс} \sin \theta \sin \omega t$ voltage from the framework during its rotation through angle θ from the direction of zero reception/procedure.

To the grids of each of the tubes $Л_1$ and $Л_2$ fall the halves of these voltages with phase displacement 180°. Amplitudes of these stresses

$$E_0 = 0{,}5 E_g \quad \text{and} \quad E = 0{,}5 E_{р\,макс} \sin \theta.$$

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Then

$$e_g = \pm (E_0 \sin \Omega t + E \sin \omega t).$$

Being limited four by terms of expansion i, for the current of the first tube, we will obtain

$$i_1 = I_0 + a_1 E_0 \sin \Omega t + a_1 E \sin \omega t + a_2 E_0^2 \sin^2 \Omega t + a_2 E^2 \sin^2 \omega t + 2a_2 E_0 E \sin \Omega t \sin \omega t + a_3 E_0^3 \sin^3 \Omega t + a_3 E^3 \sin^3 \omega t + 3a_3 E_0^2 E \sin^2 \Omega t \sin \omega t + 3a_3 E_0 E^2 \sin \Omega t \sin^2 \omega t + \dots$$

Let us isolate from this current the amplitude I_ω of component fundamental frequency ω. Current of frequency $\Omega, 2\Omega, \dots,$ as well as the high frequency harmonies 2ω, 3ω,..do not interest us since they do not pass through the high-frequency circuits connected in behind tubes Л and tuned to frequency ω.

For the current of the first tube of frequency we have

$$I_{\omega_1} = a_1 E + 2a_2 E_0 E \sin \Omega t + \frac{3}{4} a_3 E^3 + 3a_3 E_0^2 E \sin^2 \Omega t + \dots \quad (8.2)$$

Thus let us write expression for I_{ω_2} the component frequency of the current of the second tube:

$$I_{\omega_2} = -a_1 E + 2a_2 E_0 E \sin \Omega t - \frac{3}{4} a_3 E^3 - 3a_3 E_0^2 E \sin^2 \Omega t + \dots \quad (8.3)$$

Through coil L_1, occur/flow/lasts summed current $I_{\Sigma\omega}$ with frequency ω:

$$I_{\Sigma\omega}=4a_2E_0E\sin\Omega t=a_2E_{\mu}E_{p\,\text{макс}}\sin\theta\sin\Omega t. \quad (8.4)$$

During the addition of currents (8.2) and (8.3) we assume identical tubes $Л_1$ and $Л_2$ and equal of the voltage of heterodyne, supplied on both tubes. With the disturbance of these conditions, summed current will contain the unmodulated component and component, modulated by double frequency. The first of them leads to the asymmetry of the throws of the pointer of indicator from zero position during the rotation of the framework to one and another of side.

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The presence of components, modulated by the harmonics of the frequency of the reference oscillation, can cause errors. The

imbalance of the balanced modulator is undesirable for other reasons, examined below.

If I_a - radiation current and ϕ - phase displacement of the current of the balanced modulator relative to radiation current, then of emf, induced in coil at input ultrahigh-frequency (Fig. 8.3) whose inductance is further designated L_3, it will be

$$e = \omega M_2 I_a \sin \omega t + \omega M_1 I_{\Sigma\omega} \sin(\omega t + \varphi) = \\ = E_a [\sin \omega t + M \sin \Omega t \sin(\omega t + \varphi)] = E \sin(\omega t + \varphi_{рез}),$$

where M_1 and M_2 - mutual inductance L_1L_2 and L_2L_3;

$$E_a = \omega M_2 I_a; \quad M = \frac{E_p}{E_a} \sin \theta; \quad E_p = a_2 \omega M_1 E_\Omega E_{p\,макс};$$

$$\left.\begin{aligned} E &= E_a \sqrt{1 + 2M \cos\varphi \sin \Omega t + M^2 \sin^2 \Omega t}; \\ \operatorname{tg} \varphi_{рез} &= \frac{M \sin\varphi \sin \Omega t}{1 + M \cos\varphi \sin \Omega t}. \end{aligned}\right\} \quad (8.5)$$

When the radiation currents and the balanced modulator are found in phase, we obtain

$$E = E_a (1 + M \sin \Omega t). \qquad (8.6)$$

Signal is modulated in amplitude with the depth of modulation, proportional to the sine of bearing, and phase modulation is absent. After detection output potential is proportional $E_p \sin \theta$.

When ϕ = 90° amplitude of resulting emf barely depends on

bearing and modulation frequency, but the phase

$$\varphi_{\text{рез}} = \operatorname{arc\,tg}(M \sin \Omega t) \qquad (8.7)$$

changes with frequency Ω. Thus, in this case signal proves to be modulated on phase. It is known that sensitivity with phase modulation in the mode of small ratio of the voltage of signal to disturbing voltage and small index of modulation, as this is observed in radio direction finders, lower than the sensitivity with amplitude modulation.

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The structure of receiver with phase modulation is more complex than with amplitude. For these reasons in fixed loop radio compasses, is utilized the amplitude modulation, which corresponds to the absence of phase displacement between the radiation currents and the balanced modulator. However, due to inaccuracies in the production and tuning, is feasible certain phase displacement. With the phasing of the circuits of the framework and antenna, one should consider that emf, induced in them, are distinguished by phase to 90°.

Modulation factor, as can be seen from (8.5), it is proportional to sin θ. By the rotation of the framework (or by the rotation of aircraft) operator establish/installs the framework in the position,

which corresponds of zero modulation factor: $M=0$. The arrow/pointer of indicator in this case is located in the mid-position. During the divergence of the framework from zero position to one or the other side, the arrow/pointer of indicator also deviates. in one of the zero positions of rifleman/gunner's framework, it deviates the same side, as the framework. To the second zero position of the framework, which differs from the first to 180°, rifleman/gunner it deviates to the side, reverse/inverse to the framework. According to this sign/criterion it is possible to obtain one-sided bearing to radio station.

With an increase in the modulation factor, output voltage grow/rises proportionally ME_a thus far $M<1$.When begins the overmodulation $(M>1)$, stress component of the fundamental frequency F on the output of detector it is little affected with an increase in the modulation factor. At the same time grow/rise the constant component and the amplitudes of harmonics. During the rotation of the framework from the position of bearing, output current at first (when $E_p \sin\theta < E_a$) grow/rises proportional to sin θ. Subsequently beginning with the angle $\theta_0 = \arcsin\frac{E_a}{E_p}$, at which the modulation factor stops equal to unity, output current it remains almost constant (Fig. 3.4).

Since the constant component of the voltage of detector is utilized for the automatic gain control (AGC), an increase in the

constant component with the onset of overmodulation causes a fall in the amplification of receiver. During application/use AGC, the output voltage of frequency F after the onset of overmodulation does not remain constant, but begins to decrease (Fig. 8.5).

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A similar dependence impedes operator, since in some positions, revolving the framework to the side of a decrease in the throw of the pointer of indicator, it it does not approach the framework a true position of bearing, but it drives out from it. If maximum modulation factor does not exceed 1.5, a decrease in the output voltage because of action of AGC comprises not more than 2o/o, that it is possible to recognize permissible. When M=2 a decrease in the output voltage comprises already 14o/o.

The harmonics, which appear in output current with modulation, do not cause the throw of the pointer of indicator, if voltage of the generator, which feeds stator, does not contain harmonic components. But if the currents of harmonics occur/flow/last also through the stator, then the interaction of any harmonic of the output current, which takes place through the rotor, with the appropriate harmonic in armature current causes throw of pointer and, therefore, error. Therefore to also inexpediently apply large modulation factors, since

obtaining the current of the reference oscillator, free from harmonics, causes technical difficulties.

The effect of modulation factor on the sensitivity of radio direction finder is examined into §2.9.

The sources of the errors in fixed loop radio compass are different emf, induced in the channel of the framework, besides the fundamental of emf, induced within framework itself by received signal, and also modulation of the voltages in other cascade/stages, except the balanced modulator.

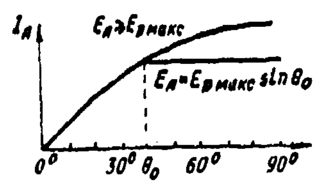

Fig. 8.4.

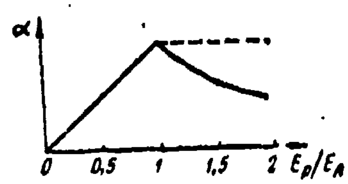

Fig 8.5

Fig. 8.4. Dependence of current in instrument on angle of rotation of framework without AGC.

Fig. 8.5. Dependence of deviation of indicator from modulation factor in the presence of AGC.

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Emf of antenna effect, which appears within the framework (§4.2), and emf, induced in it different kind by return emitters [1] (§5.3), it is possible to decompose on two components: finding in phase from tha fundamental of emf of the framework and shifted relative to it on phase to 90°.

FOOTNOTE [1]. Return emitters induce supplementary emf not only in the circuit of the framework as in the case of antenna effect, but also in the circuit of the open antenna, which leads to the onset of supplementary error from nonphase field component of reradiation, not taken into consideration (8.8) and (8.9). Supplementary error is observed independent of the accuracy of the phasing of the channels of the framework and antenna and of the tuning precision of receiver to signal. It is possible to show that the supplementary error is proportional to the square of the relation of the amplitudes of the field of reradiation and ground field and usually weight is small. ENDFOOTNOTE.

The first of them causes the same errors as in direction finder with the installation of the minimum on audition. The modulated in the

balanced modulator component, out of phase to 90°, after addition from emf of antenna creates phase modulation in accordance with (8.7). In the accurately inclined fixed loop radio compass this phase modulation does not affect indicator.

During inaccurately the channel checkup of the framework, phase displacement φ can prove to be different from zero. As a result of phase displacement in the channel of the framework, the nonphase component of antenna effect differs from emf of antenna (Fig. 8.6) not by 90°, but at an angle of 90° + φ it contains cophasal from emf of antenna component $E_{\text{н}} \sin\varphi$. where $E_{\text{н}}$ are emf of the nonphase component of antenna effect at the point of addition. Modulation factor becomes equal to

$$M=\frac{E_{\text{п}}}{E_{\text{а}}}\cos\varphi\sin\theta-\frac{E_{\text{н}}}{E_{\text{а}}}\sin\varphi.$$

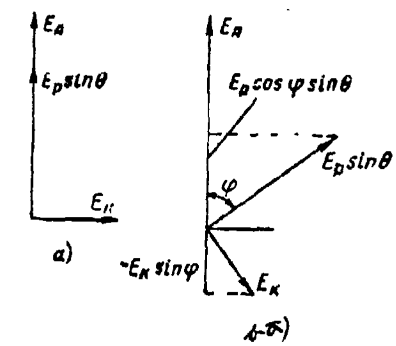

Fig. 8.6. Vector failure diagram under the effect of the extra-phase antenna effect: a) with a precise phasing of the channel of the framework; b) in the presence of phase displacement in the channel of the framework.

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Modulation factor and together with it the throw of the pointer of

indicator they are turned into zero not with $\theta = 0$, but when $\theta = \Delta_1$. Error is determined from expression $M=0$, i.e.,

$$\sin \Delta_1 = \frac{E_\kappa}{E_p} \operatorname{tg} \varphi. \qquad (8.8)$$

If $\frac{E_\kappa}{E_p} = 0{,}1$, then phase displacement must not exceed 20° so that the error will be less than 2°. The phase modulated signal, caused by antenna effect, can lead to errors also with the detuning of receiver relative to signal frequency. Due to detuning phase modulation is converted into amplitude on the ramps of resonance curve with further detection by usual detector. The output current of detector contains component modulation frequencies, caused by antenna effect, over that component which is due to the normal action of frame emf. Both these components compensate for each other, and full current is equal to zero in the position of the framework, different from the position, which corresponds to bearing, to the angle of bearing error Δ_2. Error Δ_2 is proportional to relative value of nonphase antenna effect, to modulation frequency F and to the slope of curve of resonance f'(Δf) at this detuning Δf:

$$\Delta_2 = \frac{E_\kappa}{E_p} \frac{F}{B} \frac{Bf'(\Delta f)}{f(\Delta f)}. \qquad (8.9)$$

The slope of curve of resonance depends on the degree of detuning and structure of selective system.

Table 8.1. Relative slope of curve of resonance on the band edge of transmission.

(1) Избирательная система	(2) Слабо связанные контуры				(3) Два контура при критической связи
	1	2	3	4	
Относительная крутизна $\frac{Bf'(\Delta f)}{f(\Delta f)}$	0,5	0,59	0,63	0,69	0,65

Key: (1). Selective system. (2). Weakly coupled circuits. (3). Two Ducts in critical coupling. (4). Relative slope/transconductance.

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Table 8.1 gives corrected values of the relative slope/transconductance of resonance curve on the band edge of transmission B, i.e., when $\Delta f = 0.5B$, for different selective systems.

If we on band edge allow error 1.5° when $\frac{E_к}{E_р}=0{,}12$, then passband must be $B > 3F$.

Another source of the supplementary of emf in the channel of the framework are the direct/straight communication/connections between antenna circuit and the channel of the framework. In their action they are completely analogous to antenna effect. The hook ups of antenna with the channel of the framework bear chiefly capacitive character and induce on the grids of tubes $Л_1$ and $Л_2$ co-phased voltages. For the determination of the output effect of cophasal grid voltages of the tubes of the balanced modulator, it is possible to use equations (8.2) and (8.3), changing in any of them signs for reverse/inverse and understanding by E the value of parasitic emf. During the addition of currents, the terms $2a_2E_0E \sin \Omega t$ mutually are compensated and summed current does not contain by the component modulated fundamental modulation frequency. Complete compensation occurs with equality in the absolute value of currents $I_{\omega 1}$ and $I_{\omega 2}$. During the imbalance, caused by the dissimilarity of tubes (different coefficients a_2 for tubes $Л_1$ and $Л_2$) or by the inequality of the voltages of low frequency, subjects on both tubes, the compensation is disrupted and output current contains component modulation frequencies, which causes the errors in readings of indicator. Version mentioned above second of the switching on of the tubes of the balanced modulator when voltages from the framework are fed to both grids cophasally, worse than the first version in the relation to effect the induced by antenna voltages, since parasitic voltages operate in it just as the voltage of the framework.

Very powerful signals are accompanied by errors, since the overloading of the cascade/stages of receiver causes the appearance of harmonics of modulation frequency. The latter interact with the harmonics of the current of the reference oscillation and can cause supplementary bearing error.

As indicator it is possible to apply the instrument of direct current. For its use it is necessary to rectify the output voltage of receiver with the aid of balance detector.

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8.2. The automatic direction finders WITH the servomechanism (radio compasses).

In this type radio direction finders bearing is determined by automatic unit by the antenna of system to the position of bearing with the aid of servo system.

Figures 8.7 depicts the block diagram of radio direction finder

with the servomechanism. Voltage from with the antenna of system is fed to receiver. The output voltage of receiver with the aid of control circuit controls the motor which establish/installs antenna system in the position of bearing. The output voltages of receiver must satisfy the following conditions:

1) stress must be equal to zero in the position of bearing, since motor in this position must remain motionless;

2) during deviation by the antenna of system from the position of bearing to one or another side output voltages must reverse the sign (or phase) so that the motor rotates to the proper side, i.e., in the direction, returning antenna system to the position of bearing.

The placed conditions they satisfy radio direction finders with the reading of bearing on the minimum of the depth of modulation. This direction-finding method is simpler than others, it makes it possible to carry out an automatic unit of antenna to the position of bearing. Radio direction finders with the automatic unit of the framework into the position of bearing, the operating from method determinations of minimum (zero) of modulation with cosinusoidal radiation pattern, they are called radio compasses. The field of application of radio compasses the same as fixed loop radio compasses, aircraft radio navigation.

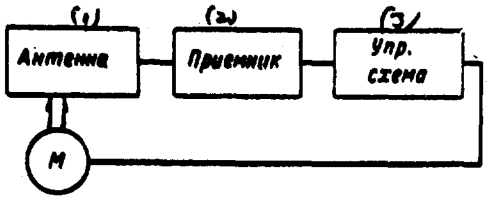

Fig. 8.7. Block diagram of radio direction finder with servomechanism.

Key: (1). Antenna. (2). Receiver. (3). Cont. diagram.

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Their fundamental operating advantage is that the operator is free/released from the search by hand for the position of bearing and from supplementary operations regarding side. Furthermore, is realized continuously the indication of the value of bearing during his changes. The antenna system of radio compass can be referred up to certain distance from operator, and this considerably facilitates the location of radio direction finder for the objective, especially on aircraft.

Antenna systems and the receptors of radio compass and fixed

loop radio compass are analogous. Output receiver current which in fixed loop radio compass supplies the indicator of bearing, in radio compass is utilized for roll control of motor. The power of output receiver currents is insufficient for the direct rotation of motor. For power gain, are applied amplifiers: tube, thyratron, magnetic, etc. Figures 8.8 depicts the widespread control circuit with the magnetic amplifiers M_1 and M_2. Since control of magnetic amplifiers is conducted by direct current, between the receiver and magnetic amplifiers is placed balance detector (tubes $Л_1$ and $Л_2$), the rectifying output receiver current with the preservation/retention/maintaining of sign with respect to the phase output current.

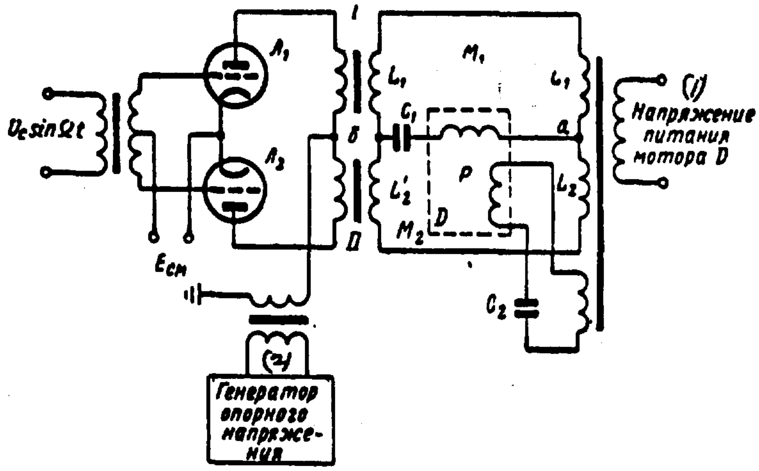

Fig. 8.8. Control circuit of radio compass.

Key: (1). Supply voltage of motor. (2). Reference generator.

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The output voltage of the receiver of the frequency of local modulation $U_c\sin\Omega t$ is fed to the grids of tubes $Л_1$ and $Л_2$

antiphase. To their anodes is fed the co-phased voltage of the same frequency from the reference oscillation. Tubes work in the mode/conditions of anode rectification on the lower bend of characteristic (grid bias E_{cM}) If the output voltage of receiver is equal to zero, the anode currents of both tubes are identical. Is identical the value of the magnetization of the arms of magnetic amplifiers I and II. If output voltage not is equal to zero and its phase coincides with the phase of reference voltage on the anode of tube $Л_1$, its anode current, that takes place on the control winding of the magnetic amplifier M_1, grow/rises, but the anode current $Л_2$, which takes place through winding M_2, falls. With respect to this and the magnetization of core I is greater than the magnetization of core II. With an alternation in the phase of output current, the magnetization of arm II becomes more than arm I.

Magnetic amplifiers are included according to bridge circuit. Bridge is comprised from inductance L'_1 and L'_2 of magnetic amplifiers and inductance L_1 and L_2 of secondary windings of the feeding transformer. Feed is conducted from the grid/network of alternating current. Output voltage is remove/taken from the diagonal of bridge ab. During the identical magnetization of the cores of both magnetic amplifiers, the bridge is balanced and voltage on diagonal ab is equal to zero. When under the effect of signal one of the cores (for example, I) it is magnetized more powerful than another, the

inductance of winding L'_1 falls, current in it grow/rises, bridge balance is disrupted and the voltage at point a becomes above than at point b. In load flows the current from a to b. During the more powerful magnetization of another core, the current in load flows to reverse side from one b to next. Thus, with an alternation in the phase of output receiver current changes the phase of the current, which feeds motor D.

As motor for the rotation of the framework, is is commonly used two-phase induction motor. One of its windings is supplied from network of alternating current, the second, control winding, by the output voltage of magnetic amplifiers from diagonal ab. Currents in supply-line and control windings must be out of phase to 90°, which is achieved by the selection of capacitors C_1 and C_2.

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Then appears the revolving magnetic field, which carries along short-circuited rotor P. With an alternation in the phase of current in control winding to 180° phase displacement of its relative to current in supply-line winding it stops - 90° magnetic field and rotor P rotate to reverse side. The rate of the rotation of motor is high. For its decrease between the motor and the framework, is placed the reducer.

Engine torque is proportional to current in control winding, who is proportional to the output voltage of receiver, i.e., to the sine of the angle of deflection of the framework from the position of bearing. Because of friction in the cell/elements of system (in motor, the bearings of the framework, reducer) the motor begins to rotate only if its torsional moment will exceed friction moment and, therefore, when the deflection of the framework will be the more than determined minimum angle, called the angle of insensitivity, or dead angle. Changes in the bearing less than this angle, they are not mastered by servo system. Dead angle must be small in comparison with the permissible error of system. Decreases in the dead angle reach

because of the use of a system with minimum friction moment and the selection of sufficient horsepower.

The most important parameters of servo system are its aperiodicity, set-up time, the speed of final adjustment and dynamic error.

Servo system must be aperiodic. During the deflection of the framework from the position of bearing, the system must return it to the position of bearing without oscillations, i.e., without the transitions through the position of equilibrium to one and another side. For providing the aperiodicity, must have the proper value braking couple on the shaft of engine, for which are utilized the supplementary hindering devices. Their braking couple must be proportional to the rotational speed. Under this condition it does not affect sensitivity, since in rest position its action does not manifest itself.

The time of establishment he is called time interval from the torque/moment of the deflection of the framework from the position of bearing to the torque/moment when servo system returns the framework to position of equilibrium. Set-up time determines the minimum duration of signal, by which can be undertaken the bearing.

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Periodic voltages, for example stress component of noises, which operate on servo system, are not caused the noticeable motion of motor, if their period is shorter than the set-up time. The band of the frequencies of the transmission of system inversely proportional to set-up time. The value of set-up time is usually 5-15 s.

The speed of final adjustment is called that maximum angular rate of rotation of the framework $\Omega_{\text{праm}}=\frac{d\theta}{dt}$, which is provided by servo system. This value has a value during a continuous change in the bearing, for example the bearing of motionless station in flight of aircraft. The usual angular rates are small, and obtaining a sufficient rate of final adjustment does not represent difficulties.

Dynamic error arises during a continuous change in the bearing as a result of the time lag of servo system. This error must be small in comparison with the common/general/total error of radio direction finder.

The methods of the calculation of control systems are the object/subject of special discipline [2.4, 2.5] and here they are not brought.

For the transmission of the position of the framework, is applied the remote system: flexible shaft, selsyn transmission, etc. Operator reads bearing on repeater. The transmission system of angle must not introduce errors.

Radio compass has two positions of equilibrium in accordance with two zeros in the radiation pattern of the framework. One of these positions stable: during a change in it at the output of receiver, appears the voltage, which sets to motion servo system to the side of a decrease in the deflection. To the second position of equilibrium, a change in the framework causes the reverse on phase voltage, which sets servo system to motion to the side of an increase in the deflection. Thus, the second position of equilibrium is unstable. Radio compass gives one-sided bearing.

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Reasons for the errors in radio compass the same as in fixed loop radio compass. Besides, that, are possible the dynamic error of servo system and the error of the system of the teletransmission of angle. These two components of errors can be made very small.

In radio compasses is applied a modulation factor less than unit, since overmodulation is unfavorable for the work of servo

system due to the appearance of sections with transimpedance on the characteristics of the dependence of the output voltage of receiver from the angle of rotation of the framework (Fig. 8.5).

Besides antennas with cosinusoidal radiation pattern, as in fixed loop radio compasses, in radio compasses it is possible to utilize the pencil-beam antennas.

The construction of circuit and in this case is most convenient during use for the direction finding of the method of determining the minimum of the depth of modulation. With the pencil-beam antennas are created two those who were displaced to certain angle of the radiation pattern whose commutation forms modulated voltage with the depth of modulation, determined by deflection from the position of bearing (see Fig. 2.15).

As a result of the sharpness of radiation pattern, is feasible the reception only in the sector, equal to the aperture angle of diagram. If is necessary direction finding from any directions, provide for two mode/conditions - search mode with long running by the antenna of system and the mode/conditions of tracking in sector.

Radio direction finders of the type in question are applied mainly at the superhigh frequencies for which easily are fulfilled

the antennas with acute/sharp radiation pattern.

8.3. Two-channel automatic direction finder with visible reading of bearing.

Figures 8.9 depicts the block diagram of this radio direction finder. Voltages from two mutually perpendicular framework or the pairs of the spaced antennas are connected to the input of receiving indicator. Receiving indicator consists of two radio receiving equipment. Figure depicts the receivers according to superheterodyne circuit, which have common/general/total heterodynes and tuned by one knob/stick.

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The output voltages of the IF amplifiers of receivers are fed to the deflector plates of cathode-ray tube - the indicator of bearings.

In radio direction finder can be used goniometric antenna system from a of the spaced antennas. Then receivers are connected to by antenna to the system through the coordinate transformer (goniometer with two mutually perpendicular search coils).

In correct work with the antenna of the system of voltage on the inputs of receivers, they will be

$$e_{10} = \mathrm{E}p_{\theta} \cos\theta \sin\omega t = E_m \cos\theta \sin\omega t,$$
$$e_{20} = \mathrm{E}p_{\theta} \sin\theta \sin\omega t = E_m \sin\theta \sin\omega t,$$

where E is strength of the field of the oriented radio station; p_{θ} — is effectiveness by the antenna of system.

Let us designate by k_1, k_2 the factors of amplification of the receivers:

$$\dot{k}_1 = k_1 e^{-j\varphi_1} \text{ and } \dot{k}_2 = k_2 e^{-j\varphi_2},$$

where k_1, k_2 are module/moduli of amplification factors;

ϕ_1, ϕ_2 - the phase shifts of the voltages in the circuits of receivers.

The output voltage of the first receiver e_{11}, usually called uptake, is connected to the vertical deflector plates of cathode-ray tube, moreover

$$e_{11} = k_1 E_m \cos\theta \sin(\omega_{\text{пр}} t - \varphi_1),$$

where $\omega_{\text{пр}}$ — is an intermediate frequency of receivers.

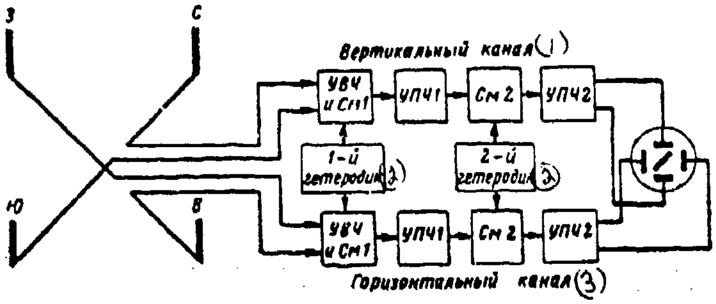

Fig. 8.9. Block diagram of two-channel radio direction finder.

Key: (1). Uptake. (2). heterodyne. (3). Tangential channel.

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The output voltage of the second receiver e_{21}, called horizontal channel, is connected to the horizontal plates of tube and

$$e_{21} = k_2 E_m \sin\theta \sin(\omega_{\text{пр}} t - \varphi_2).$$

Let in the ideal case the receivers be absolutely identical, i.e., $k_1 = k_2 = k_0$ and $\varphi_1 = \varphi_2 = \varphi_0$, and the sensitivity of tube by vertical and horizontal plates is differing and equal to k. Then the trajectory of the motion of electron beam along tube face will be straight line, inclined relative to the axis of symmetry in vertical plates to the angle, determined from the expression

$$\operatorname{tg}\alpha = \frac{kk_0 E_m \sin\theta \sin(\omega_{\text{пр}} t - \varphi_0)}{kk_0 E_m \cos\theta \sin(\omega_{\text{пр}} t - \varphi_0)} = \operatorname{tg}\theta, \ \alpha = \theta. \quad (8.10)$$

Placing on the tube a scale with divisions 0-360° so that divisions 0

and 180° will hit to the axis of the symmetry of tube, passing through vertical plates, on the glowing strip on screen, it is possible to count off two-place radio bearing. (Methods of determining the side will be presented lower). The length of the glowing strip is proportional to the strength of the field of the oriented radio station.

The dissimilarity of the module/moduli of the factors of amplification of receivers and the appearance of a phase difference of the output voltages of receivers in two-channel receiving indicator affects so, as dissimilarity of effective height and the presence of a phase difference is such of the framework of goniometric system (see § 4.6).

Let the module/moduli of the factors of amplification of receivers be equal, i.e., $k_2/k_1 = a \neq 1$, but a phase difference is absent. Then the image of bearing on the screen of cathode-ray tube is obtained in the form of the line the angle of location of which differs from azimuth by radio station by the value, determined by formula (4.16)

$$\operatorname{tg}\Delta_1 = \frac{\frac{a-1}{a+1}\sin 2\theta}{1-\frac{a-1}{a+1}\cos 2\theta}.$$

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The maximum value of error $\Delta_{1\,\text{макс}}$ is determined from expression (4.18):

$$\operatorname{tg}\Delta_{1\,\text{макс}}=\frac{\frac{a-1}{a+1}}{\sqrt{1-\left(\frac{a-1}{a+1}\right)^{2}}}.$$

The direction $\theta_{1\,\text{макс}}$, by which the error has maximum value $\Delta_{1\,\text{макс}}$, is determined from the expression

$$\theta_{1\,\text{макс}}=\frac{1}{2}\arccos\frac{a-1}{a+1}.$$

For values a, close to 1, the error has quadratic character, moreover

$$\Delta_1\approx\frac{a-1}{a+1}\sin 2\theta\approx\frac{1}{2}(a-1)\sin 2\theta \text{ and } \Delta_{1\,\text{макс}}\approx\frac{a-1}{2}.$$

During the considerable deflection of a from 1 quadratic law is disrupted. From (4.18) it follows so that the error would be not more than 0.5°, is admissible a difference in the amplification of receivers not more than approximately 2o/o.

Let the module/moduli of the factors of amplification of receivers be identical, but the output voltages of receivers are distinguished by the phase:

$$e_{11}=E_{m_1}\cos\theta\sin\omega_{\text{пр}}t,$$
$$e_{21}=E_{m_1}\sin\theta\sin(\omega_{\text{пр}}t-\varphi),$$

where $E_{m_1}=Ep_0k_0$;

φ - a phase difference between the output voltages of receivers.

In this case on tube face, is obtained the ellipse (Fig. 8.10) whose equation in polar coordinates has an expression (III.2).

The angle of the slope of the transverse of the relatively vertical axis of tube is determined by formula (4.9)

$$\operatorname{tg} 2\alpha_{\text{мин}} = \operatorname{tg} 2\theta \cos \varphi.$$

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If bearing is is counted off along the transverse than is obtained the error $\Delta_2 = \alpha - \theta$ which is designed from formula (4.20)

$$\operatorname{tg} 2\Delta_2 = -\frac{\operatorname{tg}^2 \frac{\varphi}{2} \sin 4\theta}{1 - \operatorname{tg}^2 \frac{\varphi}{2} \cos 4\theta}. \qquad (8.11)$$

For small Δ_2 the value of error can be determined by the formula

$$\Delta_2 \approx -\frac{1}{2} \frac{\operatorname{tg}^2 \frac{\varphi}{2} \sin 4\theta}{1 - \operatorname{tg}^2 \frac{\varphi}{2} \cos 4\theta}. \qquad (8.12)$$

The maximum error $\Delta_{2\text{макс}}$ and the angle $\theta_{2\text{макс}}$, which corresponds to this error, are determined by the formulas

$$|\Delta_{2\text{макс}}| = \frac{1}{4} \frac{1 - \cos \varphi}{\sqrt{\cos \varphi}},$$

$$\theta_{2\text{макс}} = \frac{1}{4} \arccos\left(\operatorname{tg}^2 \frac{\varphi}{2}\right).$$

The ratio of the minor axis of ellipse to large has expression (4.21)

$$\frac{A^2}{B^2} = \frac{1 - \sqrt{1 - \sin^2 2\theta \sin^2 \varphi}}{1 + \sqrt{1 - \sin^2 2\theta \sin^2 \varphi}}.$$

At low value φ, counting $\sin \varphi \approx \varphi$, $\sqrt{\cos \varphi} = 1$, we obtain

$$\Delta_2 \approx -\frac{\varphi^2}{8} \sin 4\theta, \quad |\Delta_{2\text{макс}}| = \frac{1}{8} \varphi^2 \text{ rad for } \theta = 22.5^\circ,$$

$$\left(\frac{A}{B}\right)^2 = \frac{\varphi^2}{4} \sin^2 2\theta, \quad \left(\frac{A}{B}\right)_{\text{макс}} = \frac{\varphi}{2} \text{ rad for } \theta = 45^\circ.$$

With small ϕ the error is octant, with an increase ϕ is developed bioctant component (law sin 8θ).

For obtaining accuracy of reading of approximately 1°, is permissible a phase difference of output voltages not more than 20° (in this case $A/B \approx 1/6$).

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When ϕ = 90° equation of ellipse (III.2) is converted into canonical:

$$\frac{x^2}{U_2^2}+\frac{y^2}{U_1^2}=1.$$

In this case with any θ the axes of ellipse coincide with the coordinate axes and direction finding is impossible.

It should be noted that with the reading of bearing on ellipse increases the subjective error of reading. It can be decreased (but it is not removed) by the application/use of the movable sight whose line operator attempts to establish/install so that it had been seemingly principal axis of ellipse.

The effect of joint action of the dissimilarity of the amplification of receivers and presence of a phase difference of the output voltages will be the same as in case examined earlier general

of the dissimilarity of the resistor/resistance of the framework in value and on phase (§4.6).

The conclusion/derivations, obtained into §4.6 in the examination of the communication/connection between the framework and the field coils of goniometer in goniometric system of two framework, are used to two-channel receiving indicator.

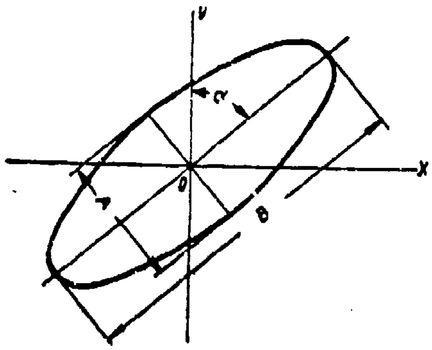

Fig. 8.10. Elliptical display.

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Visual selectivity.

Two-channel reception indicator devices with cathode-ray tube as indicator, under the condition of the linearity of processes in receiving circuits, possess the important property which calls visual selectivity. The essence of this property consists in the following. With the incidence/impingement of two signals with frequencies of f_1 and f_2 into the passband of receivers, is feasible the reading of bearings to both radio stations in their simultaneous work, if the strength of the field of signals is such, that they give

commensurable traces. Let us examine first the physical processes, which occur on scope. Let us designate $f_1 - f_2 = \Delta f$.

Under the effect only of first variation of frequency f_1 and under the condition of the identity of channels for amplification and phase blurs, of bearing on screen will be in the form of straight line ab (Fig. 8.11a). Action only of signal with a frequency of f_2 gives the bearing, determined by direct/straight CD. During the combined action of signals with frequencies of f_1 and f_2 which store/add up according to the principle of superposition, electron beam describes complex trajectory. The addition of trajectories with different frequencies can be replaced with the addition of equifrequent trajectories, which have the phase difference, which is changed in time from 0 to 360° with period $1/\Delta f$.

If a phase difference of voltage with frequencies of f_1 and f_2 changes not continuously, but irregularly after each $\Delta t = 1/f_1$ s on $(360 \cdot \Delta f/f_1)$, then for the discrete interval of time Δt on screen are drawn ellipses with different by the relation of semi-axes and by orientation.

If a phase difference is equal to 0 or 180°, the resulting trajectory will be in the form of line ~~Mm~~ MN or PL. During a change in the phase difference from 0 to 90° and from 180 to 270° ellipticity

increases, while during a change in the phase difference from 90 to 180° and from 270° to 360° - it decreases. Simultaneously with a change in the form of ellipse during a change in the phase difference is changed the orientation of its major axis from position of Mn to PL and conversely.

During the period of the beatings of the frequencies of the radio stations $1/\Delta f$, i.e., for time of a change in the phase difference from 0 to 360°, occurs the complete cycle of the transforms of images on screen (into one and another side).

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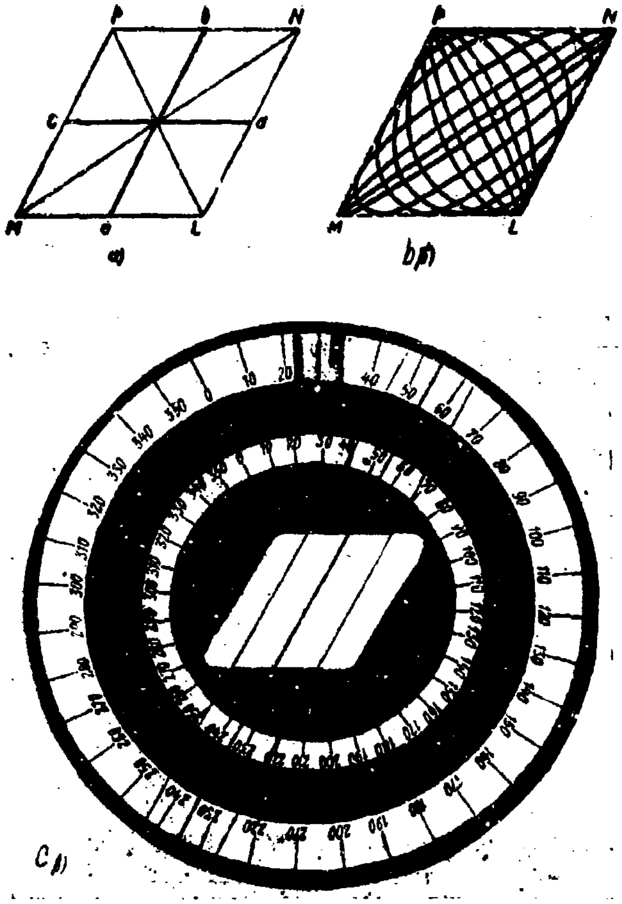

Fig. 8.11. Image of bearings of two radio stations: a) general view of parallelogram and bearings of first and second radio stations; b) exemplary/approximate form of trajectory of ray/beam and formation/education of parallelogram for time $t_{\text{вм}}=\frac{0.5}{\Delta f}$; c) parallelogram of bearings on screen of cathode-ray tube.

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The envelope of all motions of electron beam along screen forms the parallelogram whose sides are parallel to lines of bearing on radio station with frequencies of f_1 and f_2, and diagonals are lines Mn and PL.

Since in actuality a phase difference changes continuously in time, the trajectory of ray/beam will be continuously moved spiral whose component/link is formed during the period of frequency Δf and it takes the form of ellipse with the continuously changing parameters (orientation and the relationship/ratio of axes). The exemplary/approximate form of trajectory for time of a change in the phases from 0 to 180° is shown in Fig. 8.11b. The number of drawn ellipses in parallelogram is usually great.

For example for Δf = 1 kHz and f_1 = 50 kHz the number of ellipses will be $f_1/\Delta f$ = 50. Therefore separate ellipses are imperceptible, on screen is obtained the lit parallelogram.

Analytical solution to problem and obtaining the equations of the enveloping component/links of spiral for the target/purpose of simplification are carried out for the case when bearing of one of the two radio stations is 0-180° or 90-270°.

Let into the passband of receiving channels fall the signals of two radio stations in the form of sustained oscillations with frequencies ω_1 and ω_2. The bearing of the first radio station θ_1 and beam deflection on tube face along the axes OX and OY they will be

$$x_1 = U_{x1} \sin \omega_1 t,$$
$$y_1 = U_{y1} \sin \omega_1 t.$$

Here $U_{x1} = kU_{m1} \sin \theta_1$; $U_{y1} = kU_{m1} \cos \theta_1$; k - factor of proportionality (sensitivity of tube).

The bearing of the second radio station is taken equal to 0 ($\theta_2 = 0$), and the beam deflection under the effect of the output voltages, caused by the signal of the second radio station,

$$x_2 = 0,$$
$$y_2 = U_{y2} \sin \omega_2 t,$$

where $U_{y2} = kU_{m2}$.

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The combined action of the voltages of the signals of the first and second radio stations will lead to the resulting beam deflections along the axes

$$x = x_1 + x_2 = U_{x1} \sin \omega_1 t, \qquad (8.13)$$
$$y = y_1 + y_2 = U_{y1} \sin \omega_1 t + U_{y2} \sin \omega_2 t. \qquad (8.14)$$

Let us designate $\omega_2 = \omega_1 + \Delta\omega$ and convert 8.13) and (8.14). The equation of motion of electron beam will be

$$y^2 + x^2\left(\frac{U_{y1}^2}{U_{x1}^2} + \frac{U_{y2}^2}{U_{x1}^2} + 2\frac{U_{y1}U_{y2}}{U_{x1}^2}\cos\Delta\omega t\right) -$$

$$-2xy\left(\frac{U_{y1}}{U_{x1}} + \frac{U_{y2}}{U_{x1}}\cos\Delta\omega t\right) - U_{y2}^2\sin^2\Delta\omega t = 0.$$

After substituting values U_{x1}, U_{y1} and U_{y2}, we will obtain

$$y^2 + x^2\left(\operatorname{ctg}^2\theta_1 + \frac{U_{m2}^2}{U_{m1}^2\sin^2\theta_1} + 2\frac{U_{m2}\cos\theta_1}{U_{m1}\sin^2\theta_1}\cos\Delta\omega t\right) -$$

$$-2xy\left(\operatorname{ctg}\theta_1 + \frac{U_{m2}}{U_{m1}\sin\theta_1}\cos\Delta\omega t\right) - kU_{m2}^2\sin^2\Delta\omega t = 0. \quad (8.15)$$

The obtained equation is the equation of the spiral, in which continuously change the form of component/links and their orientation. The form of component/links is very close to elliptical, the frequency of a change in form and orientation of component/links is equal to $\Delta\omega$. For determining the enveloping component/links of spiral, one should solve together two systems of equations [1. 17]:

$$y(x,\ \Delta\omega t) = 0 \text{ and } \frac{\partial y(x,\ \Delta\omega t)}{\partial(\Delta\omega t)} = 0, \quad (8.16)$$

$$x(y,\ \Delta\omega t) = 0 \text{ and } \frac{\partial x(y,\ \Delta\omega t)}{\partial(\Delta\omega t)} = 0. \quad (8.17)$$

Solution for (8.16) takes the form

$$y = x\operatorname{ctg}\theta_1 \pm kU_{m2}. \quad (8.18)$$

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This equation of two sides of parallelogram, parallel the lines of bearing of the first radio station. Solution to equations (8.17) gives

$$x = \pm kU_{m1}\sin\theta_1. \quad (8.19)$$

This equation of two direct/straight, parallel axis of the ordinates, which form other two sides of parallelogram ($\theta_2 = 0$).

As already mentioned earlier, the minimum time, necessary for the formation/education of the parallelogram of visual selectivity, comprises $t_{\text{мин}} = \frac{0.5}{\Delta f}$. It should be noted that the relationship/ratio of the strengths of the fields of radio stations for time of the formation/education of parallelogram must not change. Otherwise, for example with fadings, is impeded the use of visual selectivity.

Practice shows that minimum value $\Delta f = f_2 - f_1$ with which to sufficient easily it is possible count off bearings on parallelogram on tube face without afterglow, it is approximately 10 Hz.

For the reading of bearings on parallelogram in parallel to the fundamental line of sight to the right and to the left of it they will deposit even on several lines. With direction finding are combined these lines with the sides of parallelogram, taking a reading on fundamental line.

The parallelogram of the bearings of two radio stations is shown in Fig. 8.11c.

In the case of the incidence/impingement into the passband of sensing transducers of three radio stations on screen, is observed the parallelepiped, from the sides of faces of which also are determined the bearings to all three radio stations (Fig. 8.12).

Visual selectivity is possible with the modulated in signal amplitude, but in this case the sides of parallelogram are somewhat are diffuse and the subjective errors of reading grow/rise.

The direction finding simultaneously of working radio stations substantially is facilitated, if their signals telegraph and manipulated in amplitude, since on the screen of cathode-ray tube simultaneously with parallelogram (in the case of two radio stations) or parallelepiped (in the case of three radio stations) are visible the images of the bearings of individual radio stations, which correspond thereby to the torque/moments of the time when emits one of the radio stations.

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Requirements for receiving channels.

The specific character of the work of each channel of two-channel receiving indicator is first of all determined by the facts that depending on the azimuth of signal amplitude at the inputs of channels they vary from 0 to E_m.

Consequently, both channels must have strictly linear amplitude characteristic within the limits of entire possible amplitude range of signal. Limitations on linearity begin first of all for maximum amplitudes, since can occur the overloading of final stages of one of the channels.

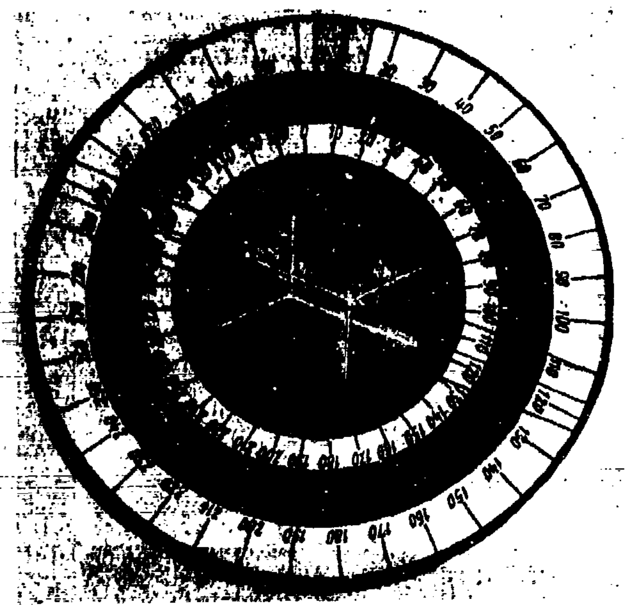

Fig. 8.12. Image of bearings of three radio stations.

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Output beyond the limits of the linear section of amplitude characteristic of one of the channels is equivalent to a change in the factor of amplification of this channel, which is led to the errors of direction finding, determined by formula (4.16).

Is feasible the case when the overloading of some cascade/stages of channel occurs under the effect of the sufficiently powerful emission/radiation of radio station, adjacent in frequency to that

which is oriented and sometimes even not giving display. Let us examine this case for amplifier and transformative cascade/stages. The module/moduli of the factors of amplification of receiving channels in general form can be presented as

$$k_1=\prod_{i=1}^{n} p_{i1}S_{i1}R_{эi1}, \qquad (8.20)$$

$$k_2=\prod_{i=1}^{n} p_{iII}S_{iII}R_{эiII}, \qquad (8.21)$$

where p_i — is the parameter, determined by the type of the resonance systems, used in receiving channel;

S_i — the slope/transconductance of the anode-grid characteristic of tube;

$R_{эi}$ — the resonance resistor/resistance of the i cascade/stage;

n - the number of cascade/stages in channel.

If one assumes that as a result of applying the different methods of the stabilization of the parameters the corresponding values $\prod_{i=1}^{n} p_{i1}R_{эi1}$ and $\prod_{i=1}^{n} p_{iII}R_{эiII}$ are made identical, then

$$a=\frac{k_1}{k_2}=\prod_{i=1}^{n}\frac{S_{iI}}{S_{iII}}.$$

Let us assume that for the assigned level of useful signal by the adjustment of the slope/transconductance of the tubes of channels it is obtained by a = 1 and the amplification of signal in channels

occurs on the linaar section of amplitude characteristic. Let on the grid of the cascade/stage of resonant amplification operate the signal $u_c = U_{mc}\sin\omega_c t$ and interference $u_n = U_{mn}\sin\omega_n t$, moreover in the presence only of signal amplifier works without overloadings.

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Let us examine the case when disturbing voltage U_{mn} is considerably greater than signal U_{mc}, but in the region of grid currents, still it does not go. Interference effect especially is strong in the first cascade/stages of receiver, since in the subsequent cascade/stages interference is attenuate/weakened as a result of selectivity. Let us examine therefore only first cascade/stages. The presence in the first cascade/stages of disturbing voltage, which envelopes into the region of grid currents, is highly improbable.

The presence of interference changes the slope/transconductance of tube for the fundamental harmonic of anode current with S_1 on [1.1, page 71]:

$$S_1 = S + \frac{1}{8} S'' U_{mc}^2 + \frac{1}{192} S'''' U_{mc}^4 + \ldots,$$

$$S_{1(n)} = S + \frac{1}{8} S'' U_{mc}^2 + \frac{1}{4} S'' U_{mn}^2 + \frac{1}{192} S'''' U_{mc}^4 + \frac{1}{32} S'''' U_{mc}^2 U_{mn}^2 + \ldots,$$

where S_1 is slope/transconductance of the tube for I of the harmonic of anode current under the effect only of signal;

$S_{1(n)}$ — the same under the effect of signal and interferences.

After considering that $U_{m c} < U_{m п}$, and after being restricted by terms of expansions with degrees U_{mc} it is not higher than the second and S'''', it is possible to obtain

$$a_{п} = \frac{S_{1(п)}}{S_1} = 1 + \frac{\frac{1}{4}\frac{S''}{S}U_{mп}^2 + \frac{1}{32}\frac{S''''}{S}U_{mc}^2 U_{mп}^2}{1 + \frac{1}{8}\frac{S''}{S}U_{mc}^2}, \qquad (8.22)$$

Further simplification in expression (8.22) we will obtain, taking into account that S''''/S is very small for the majority of tubes, a $U_{m c} \ll U_{m п}$. Then formula for $a_{п}$ will be

$$a_{п} = 1 + \delta,$$

where $\delta = \frac{1}{4}\frac{S''}{S}U_{mп}^2$.

Being given the permissible error of direction finding from overloading by interference $\Delta_{доп}$ and after determining the allowed value $\delta_{доп}$, we will obtain

$$U_{m\,п\,доп} \leqslant 2\left|\sqrt{\delta_{доп}\frac{S}{S''}}\right|. \qquad (8.23)$$

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For the case of signal in the form of sustained oscillations and interferences in the form of the modulated by tone oscillation with a modulation factor of M, it is possible to obtain

$$U_{m\,п\,доп}^{(M)} \leqslant 2\left|\sqrt{\frac{\delta_{м\,доп}}{1+2m}\frac{S}{S''}}\right|. \qquad (8.24)$$

In the case of the overloading of the mixer of inequality (8.23) and (8.24) they will be valid also, if we instead of S and S'' take conversion conductance S_{np} and its second derivative S''_{np}.

In such a way as to attenuate/weaken the effect of overloading by interference on the accuracy of direction finding, it is necessary to select operating region on the characteristic of the tube of amplifier $i_a = f(U_g)$ in area of low values S''/S and this mode/conditions of transform so that $\frac{S''_{np}}{S_{np}}$ will be minimum.

The incorrect distribution of amplification and selectivity according to receiving circuit, for example when passband sharply becomes narrow in the output stages, and the common/general/total factor of amplification of the preceding/previous cascade/stages is already sufficiently great, it can lead to the overloading of terminal and penultimate cascade/stages by the very powerful interference, considerably detuned in frequency relative to signal.

As a result of the weakening of interference by the resonance systems of the output stages, which stand after the overloaded cascade/stage, it is that commensurable with the mark of the bearing of signal smaller than its or in no way visible on screen cathode-ray

tube. This will not allow operator to note the action of interference, and the direction finding of signal will be produced with error.

Therefore during the design of two-channel receiving indicators, it is necessary to approach as far as possible to draw nearer the network elements, which determine the selectivity of receiving circuit, the input of channel, and the cascade/stages, which determine the overall gain of channel, to arrange after them.

For obtaining intensities of channels, it is necessary to have identical component values and tubes in channels.

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To ensure the maintenance of the equality of the parameters of the corresponding resonance circuits of channels is especially difficult in the circuits of hf amplification with their retuning. The errors, caused by the nonidentity of channels in high frequency, are unavoidable during a change in the tuning, since it is not possible to make a precise coupling of ducts. However, these errors can be made sufficiently small, since the passbands of high-frequency circuits are comparatively great. After using special adjustments, it is possible even more to decrease them [8.16]. Requirements for the

identity of the parameters of the resonance circuits of the narrow-band circuits of intermediate frequency are higher, but they are here facilitated by the facts that the circuits are not reconstructed in frequency [1].

FOOTNOTE [1]. Investigation of the effect of the inequality of the parameters of the resonance circuits of channels during application/use in the cascade/stages of single resonant circuits is given in [1.6, 1.9]. ENDFOOTNOTE.

Electron tubes have spread along the slope/transconductance of the anode-grid characteristics of order ±10-20o/o on linear section and even larger spreads on curvilinear sections.

Let us examine the question concerning equalization and gain control of channels by changing the grid bias voltages of tubes.

By the adjustment of bias voltage U_c and by the tuning of ducts into one of the channels it is possible to attain the equality of amplification factors in any operating mode. The identity of channels is checked using the bearing of the pilot signal which is connected to the inputs of channels with identical amplitude and phase.

In the equality of the factors of amplification of channels in amplitude and phase in scope, is observed the line at an angle of 45° to axis 0-180°.

However, due to the duration of operations on measurement and equalization of the amplification of channels, occurs the loss of the speed of the taking of bearing - the fundamental property, which characterizes two-channel radio direction finder.

Tubes have the considerable spread of the dependences

$$S_{\text{общ I}} = S_{1\,\text{I}} S_{2\,\text{I}} S_{3\,\text{I}}, \ldots, S_{n\,\text{I}} = f_1(U_{\text{см}})$$

and

$$S_{\text{общ II}} = S_{1\,\text{II}} S_{2\,\text{II}} S_{3\,\text{II}}, \ldots, S_{n\,\text{II}} = f_2(U_{\text{см}}),$$

where n - a quantity of tubes in channel;

$U_{\text{см}}$ — the bias voltage, common/general/total for all cascade/stages.

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If we consider that the parameters of the resonance systems of channels are identical, the control characteristics of the

amplification of channels can be presented in the form

$$k_1(U_{\text{см}}) = NS_{\text{общ I}}(U_{\text{см}}),$$
$$k_2(U_{\text{см}}) = NS_{\text{общ II}}(U_{\text{см}}),$$

where $N = \prod_{i=1}^{n} \mu_i R_{\text{э}\,i}$.

The exemplary/approximate course of characteristics for a two-channel receiving indicator is shown in Fig. 8.13.

The attempts to carry out a gain control of channels by one control with the preservation/retention/maintaining of the satisfactory identity of channels are led to very complex and unreliable circuits [8.16]. In this case, remanent/residual nonidentity gives errors to ±2°, ageing and the exchange of tubes they require complex original alignments, but the dynamic range of control will be small.

For the fulfillment of the automatic flare function of the amplification of channels with the possibility of gain control one by knob/stick, is applied the system in which is utilized special control signal.

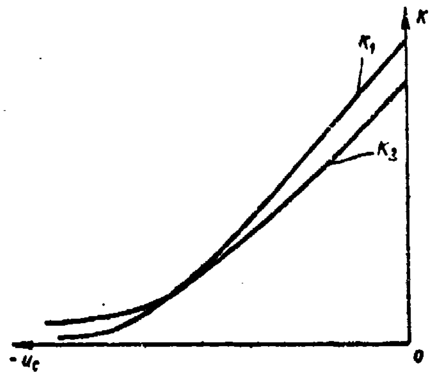

Fig. 8.13. Control characteristics of amplification of channels.

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Control signal can be form/shaped in receiving indicator itself or be formed from the oriented signal; and in that, and in the other case it must simultaneously be connected to the inputs of both channels. During the passage of control signal along receiving channels, occurs the equalization of their amplification factors. Direction finding at this time is impossible; therefore for the useful work of direction finder, control signal must be pulse with sufficiently large porosity. The frequency of filling of the radio pulse of control signal always must correspond to the resonance frequency of the tuning of receiving indicator.

Let us examine one of the principles of the formation/education

of pulse control signal in receiving indicator (Fig. 8.14). To the mixer of the control signal whose tube is triggered only during the supplying of the video pulses, which determine duration and the repetition period of steering impulses, will be feed/conducted the voltages of common/general/total for two channels first heterodyne with frequency f_{r} and special heterodyne, the generating fluctuation of the first intermediate frequency $f_{1\,\text{пр}}$.

The frequency of control signal $f_{\text{ус}}$ at the output of the mixer of control signal is equal to

$$f_{\text{ус}}=f_{\text{r}}-f_{1\,\text{пр}}.$$

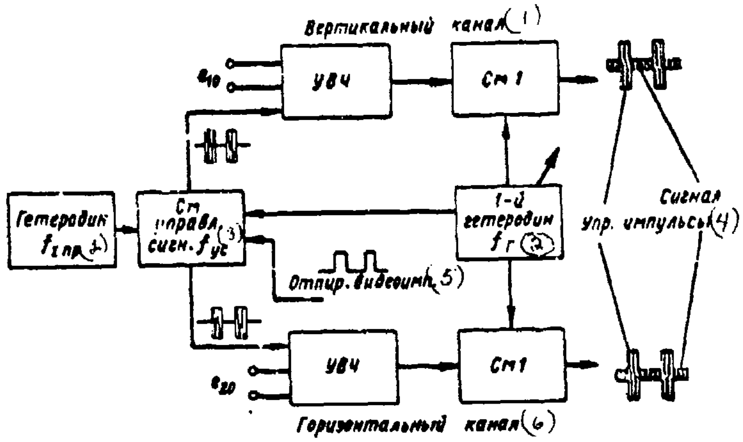

Fig. 8.14. Circuit of formation/education of pulse control signal.

Key: (1). Vertical channel. (2). Heterodyne. (3). Sm of signal control. (4). Signal. Control of pulses. (5). Triggering of videopulses. (6). Horizontal channel.

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With the reception of signal with frequency f_c and heterodyne with upper tuning $(f_r>f_c)$ in the mixers of channels is obtained the frequency

$$f_{I\,np}=f_r-f_c.$$

Then

$$f_{yc}=f_c.$$

Thus, the frequency of high-frequency filling of the momentum/impulse/pulses of control signal, changing with the retuning of receiving indicator, always it remains to the equal resonance frequency of channel checkup. The momentum/impulse/pulses of control signal are fed to the inputs of channels.

If in this case the oriented signal is not disconnected from input circuits of channels, then so that it would not affect adjustment, control signal must be several times the more than received signal.

Equalization of tae module/moduli of the factors of amplification of channels.

If $k_1 \neq k_2$, the voltages of control signal on the outputs of channels will be also different. This difference can be utilized for the equalization of the amplification of channels. The simplest circuit of the equalization of amplification of one of the channels is depicted on Fig. 8.15. This is circuit AGC with delay. The peak detector of circuit works only during the supplying of the triggering video pulses, synchronized with the momentum/impulse/pulses of control signal. Duration and the repetition period of trigger pulses are equal to duration (τ_y) and to repetition period (T_y) control signal. In time intervals $T_y - \tau_y$, when is conducted direction finding, detector is closed.

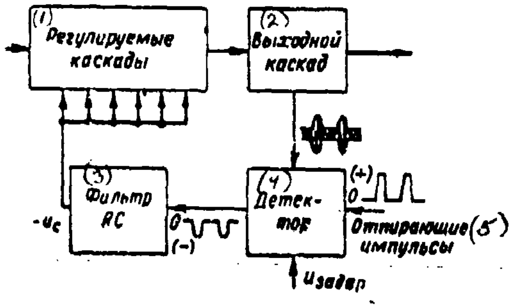

Fig. 8.15. Circuit of the equalization of the amplification of channel.

Key: (1). Adjustable cascade/stages. (2). The output stage. (3). Filter. (4). Detector. (5). Trigger pulses.

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The controlling voltage - U_c, that is obtained from detection of control signal, is fed through filter RC to control electrodes of the adjustable cascade/stages. The selection of the characteristics of the detectors, working it is separate in each channel, it makes it possible to obtain a good leveling of factors of amplification of channels. By the selection of the sufficiently slow response of the discharge of filter RC (τ_p) they attain so that the achieved/reached identity would not be disrupted in time intervals

between momentum/impulse/pulses when is conducted direction finding.

If the time constant of the discharge of capacitance/capacity C of filter does not satisfy condition $\tau_p > T_y - \tau_y$, voltage on capacitance/capacity C can considerably decrease for time of the interval/gap between steering impulses and the amplification of channels toward the end of each period T_y will increase. As a result of the possible noncoincidence of the control characteristics outside operating point, can change relation $a = k_1/k_2$ for time $T_y - \tau_y$, and this will lead to bearing error (Fig. 8.16).

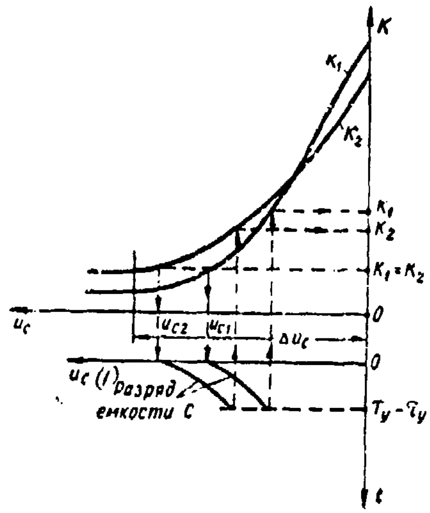

Fig. 8.16. Onset of bearing errors with the noncoincidence of control characteristics.

Key: (1). Discharge of capacitance/capacity C.

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Since each subsequent steering impulse (with is sufficient the fast time constant of the charge of capacitance/capacity C) is restored the balance of amplification, on scope with insufficient τ_p is observed the oscillation of line of bearing with frequency $\frac{1}{T_y}$ and this causes fan-shaped image (Fig. 8.17). If simultaneously is

observed the ellipticity of figure or interference, the reading of bearing becomes impossible.

Let us determine the minimally permissible value $\tau_p = RC$ at the assigned magnitude of the permissible error of direction finding, caused by the discharge of capacitance/capacity C. In this case, let us consider that the control characteristics of amplification $k_1 = f_1(U_c)$ and $k_2 = f_2(U_c)$ are uniform and are approximated by exponential dependence, i.e., $k = k_0 e^{-\alpha U_c}$, where α is average mutual conductance of control; U_c displacement on control electrodes of the adjustable tubes; k_0 is an initial value of the amplification factor when $U_c = 0$.

We take, that the time constants of the discharge of the filters of each channel are equal to each other $(\tau_{p1} = \tau_{p2} = \tau_p = RC)$ and that toward the end of the duration of each steering impulse occurs the absolute equalization of channels on the module/modulus of factor of amplification, i.e.,

$$k_{0\,I} e^{-\alpha U_{c1}} = k_{0\,II} e^{-\alpha U_{c2}}. \qquad (8.25)$$

From the torque/moment of the termination of the action of steering impulse, capacitors C in the filters of channels, charged to potentials U_{c1} and U_{c2} respectively, in time $T_y - \tau_y$ are discharged, moreover $\tau_p = RC$ (Fig. 8.16).

For the torque/moment of the beginning of the subsequent steering impulse, the factors of amplification of channels will change differently and they will be equal to

$$k_1 = k_{01} \cdot \exp\left[-\alpha_1 U_{c1} \exp\left(-\frac{T_y - \tau_y}{\tau_p}\right)\right], \quad (8.26)$$

$$k_2 = k_{011} \exp\left[-\alpha_2 U_{c2} \exp\left(-\frac{T_y - \tau_y}{\tau_p}\right)\right]. \quad (8.27)$$

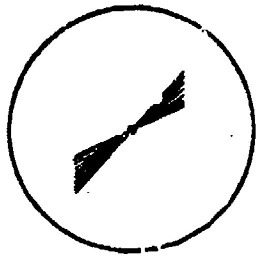

Fig. 8.17. Image of bearing with the fast time constant of the circuit of the eqialization of channels.

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Let us find the relation $a_p = \frac{k_1}{k_2}$, which determines bearinj error due to the appearance of a nonidentity of channels accordinj to amplification toward the end of each period T_y (i.e. when $t = T_y - \tau_y$).

From (8.26) and (8.27), utilizing (8.25), it is possible to determine

$$a_p = \frac{k_1}{k_2} = A_0^{\left[1-\exp\left(-\frac{T_y-\tau_y}{\tau_p}\right)\right]}, \qquad (8.28)$$

where $A_0 = \frac{k_{0\,I}}{k_{0\,II}}$.

From (8.28) it follows that a_p depends on the initial imbalance of

channels A_0 (when $U_{c_1}=U_{c_2}=0$). With A_0 = 1 errors of direction finding, even with fast time constants τ_p, will not be, if control characteristic is exponential.

Considering that $1<a_p<1{,}02$ (error of direction finding is not more than 0.5°), $A_0>1$, and $\frac{T_y-\tau_y}{\tau_p}\ll 1$, it is possible to record

$$\left.\begin{aligned}&\ln\frac{a_p}{A_0}\approx\frac{a_p}{A_0}-1,\\&\ln\frac{1}{A_0}\approx\frac{1}{A_0}-1,\\&\exp\left(-\frac{T_y-\tau_y}{\tau_p}\right)\approx 1-\frac{T_y-\tau_y}{\tau_p}.\end{aligned}\right\} \quad (8.29)$$

Utilizing (8.29), from (8.28) we will obtain

$$\tau_p>\frac{A_0-1}{a_p-1}(T_y+\tau_y).$$

For example, with A_0 = 2, $T_y+\tau_y=20$ ms and $a_p=1{,}02$ (errors of direction finding do not exceed (0.5°) the allowed value of the time constant of the discharge of the capacitance/capacity of filter will be determined

$$\tau_p\geqslant 1 \quad \text{s}.$$

Control of amplification during the use of a pulse control signal.

A change in the value of control signal at the inputs of channels will cause the appropriate change in the output voltages of control signals, which in turn, through the circuit of the equalization of amplification will lead to a change in the factors of amplification of channels.

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The level of control signal at the inputs of channels changes with the adjustment of bias voltage on the grid of the tube of the mixer of control signal.

Thus far the values of the obtained in this case voltages of control U_{c1} and U_{c2} do not exceed the limits of the range of control (ΔU_c on Fig. 8.16), simultaneously with a change in the amplification occurs its equalization. Consequently, the dynamic range of gain control will be the wider, the lesser the scatter of control characteristics.

On the other hand, the dynamic range of the control of amplification is determined by the value of the linear section of function $U_{nmx}=f(U_{nx})$, where U_{nx} and U_{nmx} voltage on

entrance and exit of channel.

The level of control signal can be regulated both by hand and is automatically (Fig. 8.18).

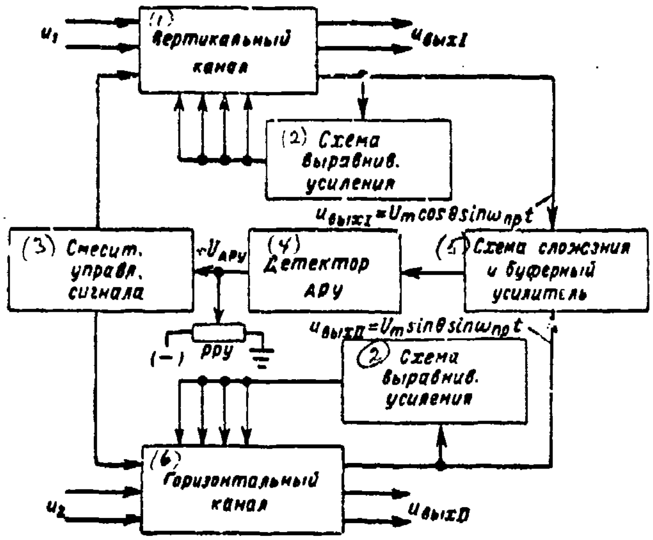

Fig. 8.18. Circuit of automatic gain control.

Key: (1). Uptake. (2). Circuit of amplification balance. (3). Mixer control of signal. (4). Detector. (5). Adding circuit and buffer amplifier. (6). Tangential channel.

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The voltages of the oriented signal in receiving channels depend on the value of bearing. With the formation of voltage for the automatic gain control this dependence must be excluded. For this, the parts of output voltages of channels store/add up in quadrature

(with phase displacement to 90°, usually realized by chain/networks RC).

The resulting voltage, equal to

$$U_{1\cdot 03}=\sqrt{U^2_{m\,вых}\cos^2\theta+U^2_{m\,вых}\sin^2\theta}=U_{m\,вых}$$

and not depending on θ, is fed to detector of AGC. Detector of AGC is included so that the unidirectional voltage of the signal of positive polarity decreases the initial negative displacement on the grid of the mixer of the control signal whose level in this case increases. Thus, AGC and the equalization of the amplification of channels are combined.

The time constant of circuit of AGC is determined by the time constant of the filter of the equalization of amplification.

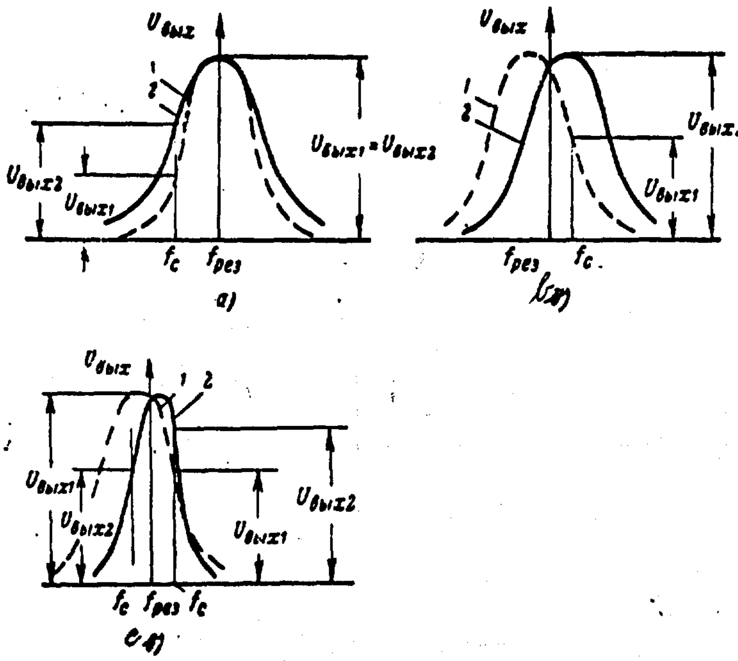

Fig. 8.19. Reasons for the oscillation of bearings with detuning. a) the different qualities of resonance systems; b) the mutual detuning of channels; c) the mutual detuning of channels and the different qualities of resonance systems; 1 - resonance curve of uptake; 2 - resonance curve of tangential channel.

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It is necessary to keep in mind that in system with the pulse control signal, the frequency of filling of which is equal to the resonance frequency of the tuning of receiving indicator, the

equalization of amplification occurs only at this frequency. Therefore, if the resonance characteristics of channels do not coincide in form friend with other within the limits of passband, during imprecise tuning to signal, can arise the errors of direction finding, caused by the nonidentity of the resonance characteristics of channels. These errors are developed in the deviation of bearing from true with detunings receiving indicator (the so-called oscillations of bearing). Depending on the reasons for the nonidentity of the resonance characteristics (different energy factor of ducts, staggering) of oscillation, they can have symmetrical and asymmetric character with detunings to both sides from resonance (Fig. 8.19).

So, in the case Fig. 8.19a the oscillation of bearing they occur to one side from the image of the true bearing; in the case Fig. 8.19b, appear symmetrical oscillations to both sides; in the case Fig. 8.19c, occur asymmetric oscillations to both sides.

Equalization of the phase shifts of the voltage in channels.

The phase shifts of high-frequency filling of steering impulses, obtained after passage by them receiving channels, are utilized for

equalization within certain limits of phase characteristics of channels.

One of the possible circuits of phase compensation it is shown to Fig. 8.20. The steering impulses through the buffer stages enter the phase discriminator. The buffer stages are open/disclosed by the video pulses, synchronous and cophasal with managers, due to what the circuit of phase compensation works only under the effect of steering impulses and does not react to the oriented signal. Manufactured by phase discriminator regulating direct/constant voltages (U_p) are fed to the circuit of control of the phase of the stress of the second heterodyne, common/general/total for two channels. This circuit consists of two phase-shifting cascade/stages, connected between the second heterodyne and the second mixers of receiving channels.

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As is known, the phase of the voltage of intermediate frequency is determined by the phase of the voltage of signal and by the phase of heterodyne $(\varphi_{np}=f(\varphi_c, \varphi_r))$. Under the effect of the controlling voltage of phase discriminator, the voltage of the second heterodyne in the phase-shifting cascade/stage acquires the phase shift, equal in magnitude to the phase shift of governing output potential of this

channel and reverse/inverse on sign. As a result of the phase of output voltages, they are equalized.

The preservation/retention/maintaining of the identity of the phases into time intervals $T_y - \tau_y$ is assured by selection sufficient slow response of the discharge of a RC-filter in the circuit of shaping of output direct/constant controlling voltages of phase discriminator.

Besides the system described above of the equalization of channels with the aid of control signal, are applied other methods [8.3].

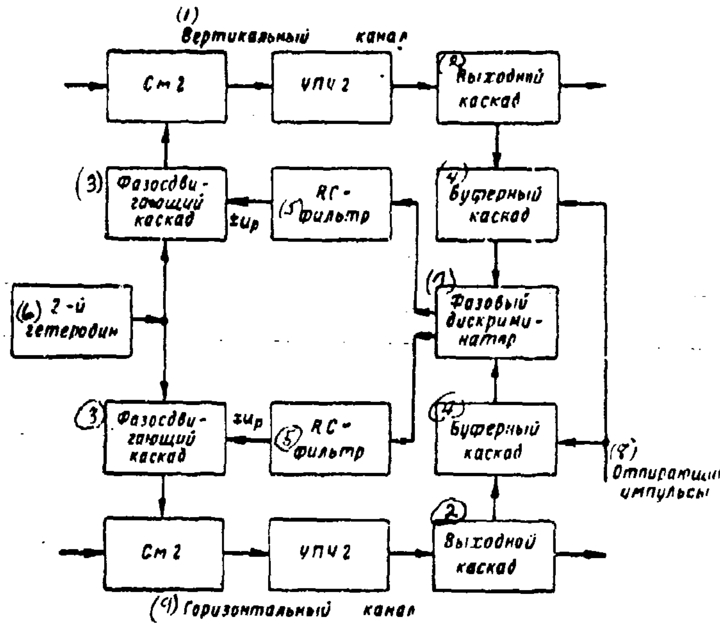

Fig. 8.20. Circuits of phase compensation.

Key: (1). Uptake. (2). The output stage. (3). Phase-shifting cascade/stage. (4). The buffer stage. (5). filter. (6). heterodyne. (7). Phase discriminator. (8). Trigger pulses. (9). Tangential channel.

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It is possible to decrease the requirements for the identity of channels with respect to amplification or at least continuous to

monitor it, without applying special pilot signal, if we at entrance and exit of receiving indicator periodically change over channels, changing by their places (Fig. 8.21) [8.17]. In this direction finder the disturbance/breakdown of the identity of channels on amplification and phase will lead to the fact that on scope they are formed two lattice-type ellipses. Correct bearing is determined from the bisector of the angle between the principal axes of ellipses (Fig. 8.22). During the equalization of amplification, the ellipses will merge into one.

Such by shape, in this system is feasible the correct reading of bearing with the sufficiently considerable nonidentity of channels and it is easy to note the disturbance/breakdown of identity. But the accuracy of reading of bearing (on bisector) descends, and when interference and instability of the figure of bearing is present, reading is extremely hinder/hampered. The minimum frequency of commutation with tube without afterglow must be order 20 Hz.

Obtaining one-sided bearing.

The two-channel radio direction finder, working according to these circuits, makes it possible to obtain bearing with

indeterminancy/uncertainty into 180°. For the elimination of ambiguity, it is possible to apply different circuits. Usually in majority of them, is utilized the superposition principle of the nondirectional reception on the directional cosinusoidal reception.

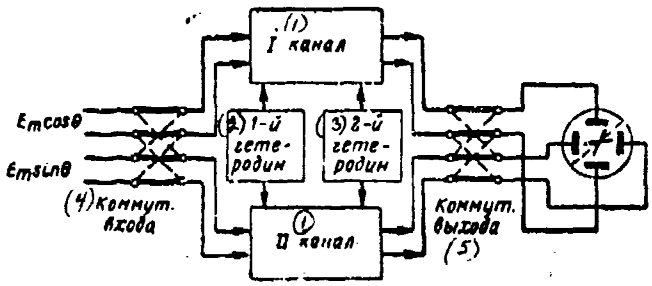

Fig. 8.21.

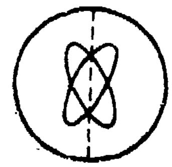

Fig. 8.22.

Fig. 8.21. Block diagram of two-channel radio direction finder with switching of channels.

Key: (1). channel. (2). the 1st heterodyne. (3). the 2nd heterodyne. (4). input switching. (5). output switching.

Fig. 8.22. Image of bearing with the nonequalized channels.

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Most convenient, that not requiring supplementary operations of operator, but also most complex on execution is the three-channel circuit of direction finder. To the input of the third channel, is connected the central omnidirectional antenna, and to its output -

the modulator grid of brightness or the focusing grid of cathode-ray tube (Fig. 8.23).

The process of obtaining the single-valued bearing it is easy to explain by diagram/curves Fig. 8.24. On diagram/curves 1 and 2, are shown the voltages of vertical and tangential channels. The relative phase of these stresses depending on the direction of the arrival of wave can change by 180°.

On diagram/curve by 3 solid lines is shown the voltage of the omnidirectional antenna, wattful of the directional antenna. The phase of the omnidirectional antenna does not change with a change in the bearing. The half-periods of the work of cathode-ray tube are shown on diagram/curve 4.

The form of line of bearing during control by the modulator of brightness for case Fig. 8.24 is shown to Fig. 8.25a. With the noncoincidence of the phases of the stresses of the omnidirectional antenna and directed system (dotted line on diagram/curve 3) (Fig. 8.24) the form of line of bearing will be such, as on Fig. 8.25b. If a phase difference reaches 90°, bearing will be two-place, with an increase its more than 90° determination of side will be incorrect.

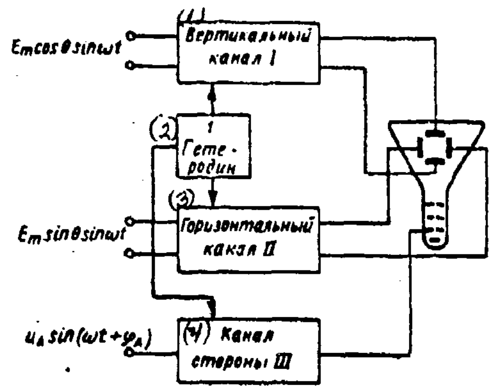

Fig. 8.23. Circuit of radio direction finder with the separate channel of the voltage amplification of the omnidirectional antenna.

Key: (1). Uptake. (2). Heterodyne. (3). Tangential channel. (4). Channel of side.

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Thus, in three-channel system it is necessary to ensure very high requirements for the identity of all channels with respect to phase shift.

A simpler pattern of the determination of side, worker on the same principle, can be realized without the third channel. The inputs

of channels alternately are connected to the omnidirectional antenna; the output of that channel which is connected to it at given torque/moment, is connected to the modulator of the brightness of tube (Fig. 8.26). Second channel remains at this torque/moment connected to its deflector plates. Such switchings can be produced by hand or after include/connecting the special switching equipment/device (in the simplest execution relay). During manual switching it is necessary to be guided by following. If bearing is close to values of 0-180°, the omnidirectional antenna is connected to the tangential channel whose output voltage will be feed/conducted in this case to the modulator of brightness. With the bearing, close to 90-270°, the channels are changed by places. The diagrams of the voltages in channels and the form of display are shown to Fig. 8.27.

If at the input of direction finder there is a goniometer, it suffices to connect to nondirectional antenna and modulator grid only one of the channels, since in this case the voltage of the directional antenna is distributed along channels by goniometer. All requirements for the phase relationship/ratios of the voltages of channels during the determination of the sides, given above, are valid also for this circuit.

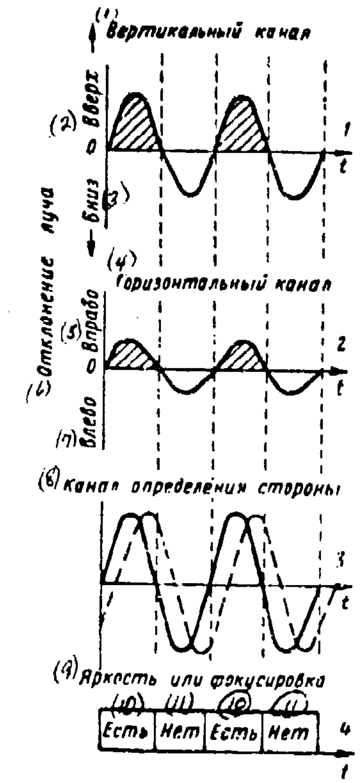

Fig. 8.24. Process of obtaining the single-valued bearing in circuit Fig. 8.23.

Key: (1). Uptake. (2). Upward. (3). Down. (4). Tangential channel. (5). To the right. (6). Beam deflection. (7). To the left. (8). Channel of the determination of side. (9). Brightness or focusing. (10). There is. (11). No.

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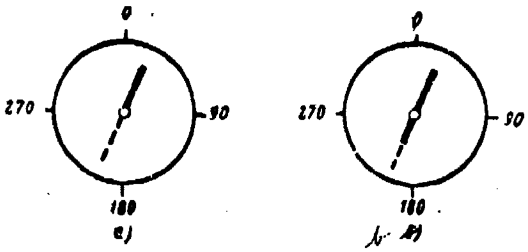

Fig. 8.25. Image of single-valued bearing.

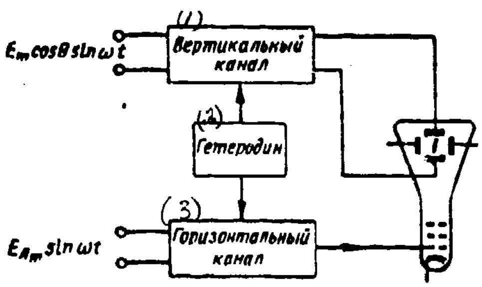

Fig. 8.26.

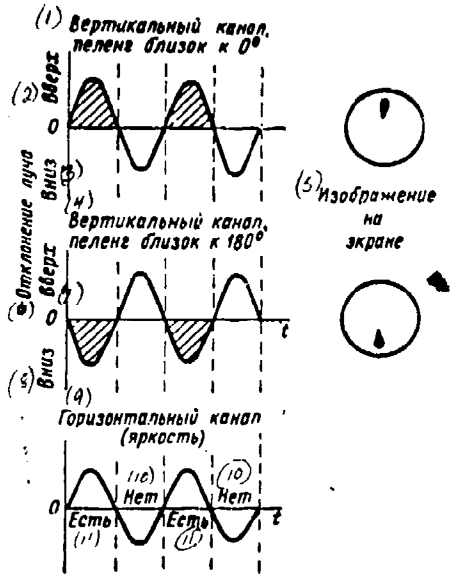

Fig. 8.27.

Fig. 8.26. Patter1 of the determination of side in two-channel RDF.

Key: (1). Uptake. (2). Heterodyne. (3). Tangential channel.

Fig. 8.27. Diagrams of voltages and image on screen of CRT for circuit Fig. 8.26.

Key: (1). Uptake, bearing are close to 0°. (2). Upward. (3). Down. (4). Uptake, bearing are close to 180°. (5). Image on screen. (6). Beam deflection. (7). Upward. (8). Down. (9). Tangential channel (brightness). (10). No. (11). There is.

8.4. Single-channel radio direction finders.

The difficulties of the execution of two identical channels of amplification led to the creation of single-channel radio direction finders. According to operating principle, they are two-channel, on physical channel separation, it is replaced in them by frequency or time/temporary. To this same to class can be attributed the radio direction finders in which amplifier circuits are united partially, channel separation occurring according to frequency sign/criterion. It should be noted that the association of amplifier circuits frees from some difficulties and the sources of the errors which are observed during two-channel amplification, but at the same time complicates circuit (modulation or commutation, etc.), are caused the new sources of errors. The majority of single-channel RDF's is

deprived of one of the fundamental advantages of two-channel radio direction finders - visual selectivity.

Direction finders with audio modulation (method of comparison of the depth of modulation of the taken signal).

The block diagram of direction finder is depicted on Fig. 8.28. The voltages of signal u_1 and u_2 will be feed/conducted to two balanced modulators and modulated by the voltages of two frequencies Ω_1 and Ω_2 respectively. The output voltages of the balanced modulators will be

$$u_{м1} = U_м \cos\theta \sin\Omega_1 t \sin\omega t,$$
$$u_{м2} = U_м \sin\theta \sin\Omega_2 t \sin\omega t.$$

After addition with the voltage of the omnidirectional antenna

$$u_а = U_{m\,а} \sin\omega t$$

on the input of receiving indicator, operates the voltage

$$u = u_a + u_{M1} + u_{M2} = U_{m\,a}\left[1 + \frac{U_M}{U_{m\,a}}\cos\theta\sin\Omega_1 t + \right.$$
$$\left. + \frac{U_M}{U_{m\,a}}\sin\theta\sin\Omega_2 t\right]\sin\omega t$$

or

$$u = U_{m\,a}\left[1 + M_1\sin\Omega_1 t + M_2\sin\Omega_2 t\right]\sin\omega t, \quad (8.30)$$

where $M_1 = \frac{U_M}{U_{m\,a}}\cos\theta;\ M_2 = \frac{U_M}{U_{m\,a}}\sin\theta.$

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From (8.30) it is evident that at the input of receiving indicator operates the voltage of carrier frequency ω, modulated in amplitude by the voltages of frequencies Ω_1 and Ω_2 with the modulation factors M_1 and M_2, which lepend on the azimuth of the DF radio station.

Such by shape, the information about bearing in this direction finder is included in the value of the side components of the modulated by two tones signal and their mutual cophasality or pushpulis (last/latter determines the quadrant in which is located the single-valued bearing).

To avoid cross modulation in this circuit, the overmodulation is not admitted and is necessary satisfaction of condition $M_1 + M_2 < 1$. It is obvious that for this it is necessary to satisfy the condition $\frac{U_м}{U_{m\,л}} < \frac{1}{\sqrt{2}}$, since the maximum value of sum cos θ + sin θ is equal to $\sqrt{2}$.

After detection the variable component of signal at the output of receiver takes the form

$$u_л = U_{m\,л} \cos \theta \sin \Omega_1 t + U_{m\,л} \sin \theta \sin \Omega_2 t. \qquad (8.31)$$

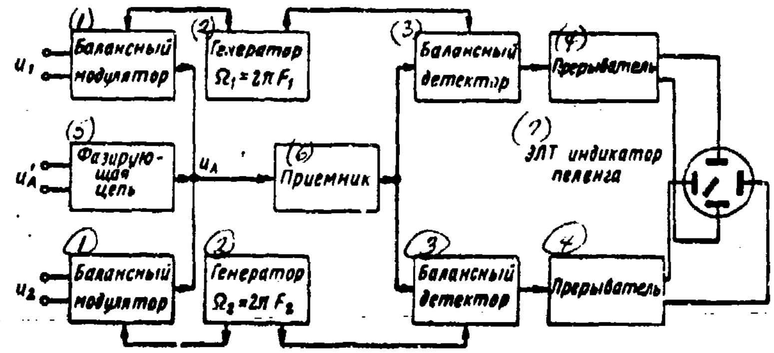

Fig. 8.28. Block diagram of direction finder with audio modulation.

Key: (1). The balanced modulator. (2). Generator. (3). Balance detector. (4). Interrupter. (5). Phasing circuit. (6). Receiver. (7). CRT bearing indicator.

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Voltage (8.31) is fed to two synchronous balance detectors to which are conducted reference voltages $u_{\Omega 1}=U_{\Omega m}\sin\Omega_1 t$ and $u_{\Omega 2}=U_{\Omega m}\sin\Omega_2 t$. During the phase coincidence of the supporting/reference and worker of voltages (or with their pushpull) the output voltages of balance detectors will be

$$U_{I}=U_{mc}\cos\theta,\qquad U_{II}=U_{mc}\sin\theta. \tag{8.32}$$

Direct/constant voltages (8.32) are utilized for the indication of bearing with the aid of dial instrument (magnitoelectric logometer) [8.12].

With the supply of voltages (8.32) to the plates of cathode-ray tube the focus will be deflected from center and its angular position will give radio station bearing. More convenient indication with the aid of cathode-ray tube is obtained, if we include/connect interrupters (Fig. 8.29), converting direct/constant voltages (8.32) in saw-tooth.

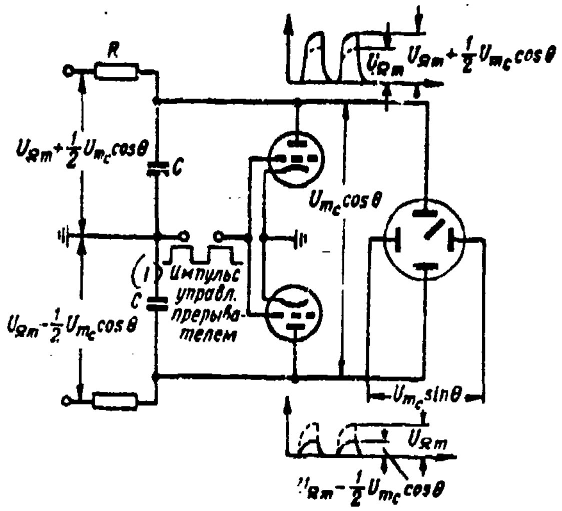

Fig. 8.29. Circuit of the formation/education of saw-tooth voltages.

Key: (1). Momentum/impulse/pulse control of interrupter.

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Image on the screen of cathode-ray tube takes in this case the form of the radial ray/beam on which is counted off single-valued bearing.

In the direction finder in question partially is retained the separate passage of the voltages of two directional antennas along independent channels. In this part of the circuit, is possible the

onset of the errors, examined into § 8.3.

The work of the balanced modulators is examined in § 8.1. The obtained there conclusion/derivations can be attributed also to this circuit.

The amplification of side frequencies $\omega \pm \Omega_1$ and $\omega \pm \Omega_2$ in common/general/total circuit must be identical. The different value of amplification factors causes the errors, analogous to the errors of two-channel radio direction finder. A difference in the envelope phases of variations of frequencies Ω_1 and Ω_2 is led to errors, since balance detector converts a phase difference into a difference in the amplitudes of output voltages. Actually, if phase displacements of envelopes of relatively reference voltage comprise $\Delta\varphi_{or1}$ and $\Delta\varphi_{or2}$, then the output voltages of balance detectors take the form

$$U_{I} = U_{m0} \cos \theta \cos \Delta\varphi_{or1},$$
$$U_{II} = U_{m0} \sin \theta \cos \Delta\varphi_{or2}.$$

The angle of deflection of ray/beam on the screen of the cathode-ray tube of indicator is determined by the expression

$$\operatorname{tg} \Phi = \frac{U_{II}}{U_{I}} = \xi \operatorname{tg} \theta,$$

where $\xi = \frac{\cos \Delta\varphi_{or2}}{\cos \Delta\varphi_{or1}}$. Bearing error (Δ_{or}) in this case is analogous to error with the inequality of the module/moduli of the factors of amplification of two-channel indicator and can be found from formula

(see § 8.3)

$$\mathrm{tg}\,\Delta_{\text{er}} = \frac{\frac{\xi - 1}{\xi + 1}\sin 2\theta}{1 - \frac{\xi - 1}{\xi + 1}\cos 2\theta}.$$

When selecting modulation frequencies, must be take into consideration of condition, analogous to the conditions for the selecting of the frequency of rotation in phase-meter radio direction finders (see § 8.5).

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Direction finders with the successive connection of antennas and indicator.

In such direction finders the orthogonal pairs of antennas or coil of goniometer alternately are connected to single-channel receiving indicator. Simultaneously are switched the coils of the needle indicator (logometer) or the deflector plates of cathode-ray tube. The simplest circuit of this direction finder is depicted on Fig. 8.30. For the stabilization of bearing, the indicator must have slow response. Direction finders with switching did not have

extensive application due to the large number of deficiency/lacks.

Two-channel reception indicator with the partial association of channels.

The difficulties of the control of the amplification of channels in systems without control signal led to the creation of two-channel receiving indicator with the partial association of channels [8.3]. In this receiving indicator is conducted the special conversion of intermediate frequency in one of the channels, so that the intermediate frequencies of channels considerably would differ from each other (Fig. 8.31). Then the signals of each channel with different intermediate frequencies are amplified on the whole, usually the aperiodic, amplifier.

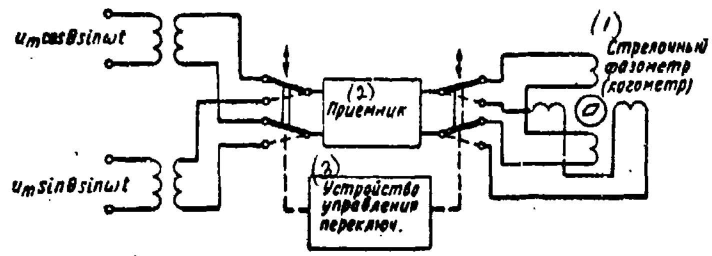

Fig. 8.30. Block diagram of single-channel radio direction finder with the successive connection of antennas and indicator.

Key: (1). Arrow phasemeter (logometer). (2). Receiver. (3). Control unit switch.

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With a sufficient broad-band character of this amplifier and linear amplitude-frequency characteristic with the adjustment of its amplification it is possible within certain limits to change to identical degree both signal level.

For obtaining the indication of bearing on cathode-ray tube it is necessary again to convert frequencies, in order to obtain them identical. Since conversion before the common/general/total amplifier compulsorily is led to a supplementary difference of the signals in

channels due to phase, for reverse/inverse transformation is utilized the same heterodyne, but connected to the mixer of another channel (Fig. 8.31).

Usually the dynamic range of the control of amplification by such circuits does not exceed 40-50 dB. The automatic gain control also can be realized in common/general/total amplifier. As in system with control signal, stress level on the input of detector of AGC must not depend on azimuth (§ 8.3).

8.5. Instrument errors of RDF's because of phase shifts in receiver.

During the application/use of methods of radio traffic, based on the measurement of phase, vital importance acquire phase shifts in receiver. Phase measurements in high frequency are utilized in phase radio direction finders from motionless antennas with large separation.

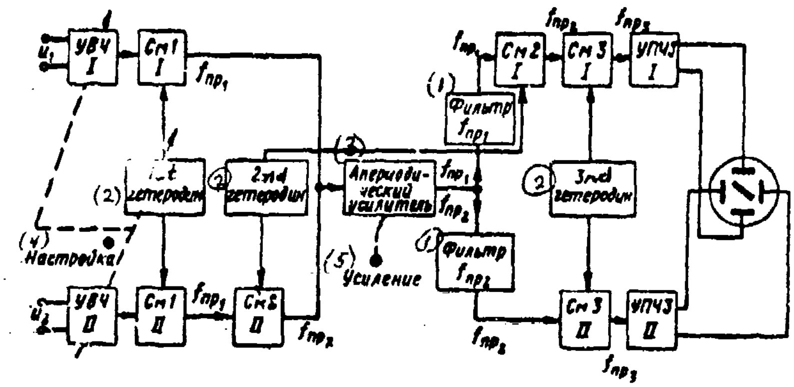

Fig. 8.31. Block diagram of receiving indicator with the partial association of channels.

Key: (1). Filter. (2). heterodyne. (3). Aperiodic amplifier. (4). Tuning. (5). Amplification.

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The measurement of envelope phase of the amplitude- or frequency modulated oscillation is applied in radio direction finders with the long running of antennas and in radio direction finders with the cyclic measurement of phase. In the enumerated systems the information about bearing is included directly in phase. Therefore any supplementary negligible phase displacement in receiving

indicator will give corresponding bearing error. At the same time the presence in the receiving indicator of resonance systems in the amplifiers of the high and intermediate of frequencies unavoidably is led to phase displacement both the carrier frequency and of envelope. An especially large shift occurs in circuit of IFA [intermediate-frequency amplifier], where the there is the narrowest passband of resonance systems and the steepest phase response.

As is known, the phase response of single duct and two coupled circuits can be expressed by the formulas:

$$\varphi_K(\sigma) = -\operatorname{arc\,tg} \sigma, \qquad (8.33)$$

and

$$\varphi(\sigma) = -\operatorname{arc\,tg} \frac{2\sigma}{1-\sigma^2+\eta^2}, \qquad (8.34)$$

where $\sigma = (2\Delta f/f_0)Q$ - the generalized detuning;

f_0 - resonance frequency;

$\Delta f = f - f_0$ - detuning;

Q - a quality;

$\eta = KQ$ - coupling parameter;

K - a coupling coefficient.

The selective system of receiver usually consists of several free single ducts or pairs of circuits. Resulting phase displacement grow/rises proportional to n - to the number of single ducts or pairs of ducts in system. If the resulting passband B is assigned, then the energy factor of ducts must be selected as being equal to

$$Q=\frac{f_0}{B}\frac{1}{\psi(n)}. \qquad (8.35)$$

Function $\psi(n)$ in the case of applying the pairs of ducts depends besides n and on the degree of communication/connection (coupling parameter η). Table 8.2 gives corrected values $\psi(n)$, borrowed from [2.8].

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Utilizing (8.35) and designating x = 2Δf/B, we obtain

$$\alpha=\frac{x}{\psi(n)}. \qquad (8.36)$$

The phase responses of multistage selective system can be, using these designations, recorded in the form

$$\varphi_n=-n\,\mathrm{arc\,tg}\frac{x}{\psi(n)}, \qquad (8.37)$$

for a single-circuit circuit, and

$$\varphi_n = -n \operatorname{arc\,tg} \frac{2x\psi(n)}{(1+\eta^2)\psi(n)^2 - x^2} \qquad (8.38)$$

for a two-circuit circuit.

The phase response of resonance system for an envelope sinusoidal oscillation, modulated voltage of frequency F, is determined by formula [2.7]

$$\varphi_{0r}(\sigma) = \frac{d\varphi(\sigma)}{d\sigma}\frac{2F}{f_0}Q. \qquad (8.39)$$

For the single duct

$$\varphi_{0r}(\sigma) = -\frac{1}{1+\sigma^2}\frac{2F}{f_0}Q; \qquad (8.40)$$

for a system of two coupled circuits

$$\varphi_{0r}(\sigma) = -2\frac{1+\sigma^2+\eta^2}{(1-\sigma^2+\eta^2)^2+4\sigma^2}\frac{2F}{f_0}Q. \qquad (8.41)$$

Table 8.2. Values

(1) Схема	(2) Режим настройки	(3) $\varphi(n)$ при числе каскадов					
		1	2	3	4	5	6
(4) Одноконтурная	(5) Все контуры в резонансе	1,0	1,56	1,96	2,3	2,85	2,89
(6) Двухконтурная	$\eta = 0{,}5$	1,19	1,71	2,06	2,35	2,67	2,94
	$\eta = 1$	0,71	0,88	0,98	1,09	1,16	1,22
	$\eta = \eta_{макс}$	0,32	0,46	0,55	0,61	0,67	0,7

Key: (1). Circuit. (2). Mode/conditions of tuning. (3). $\varphi(n)$ with the number of cascade/stages. (4). Single-circuit. (5). All ducts in resonance. (6). Two-circuit.

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As can be seen from (8.41), the form of curve $\varphi_{or}(\sigma)$ depends on η. So that curve $\varphi_{or}(\sigma)$ would have one maximum, is necessary satisfaction of condition $\eta < \frac{1}{\sqrt{3}}$.

Phase displacement of envelope is maximum during the fine tuning of resonance system for signal carrier frequency (σ = 0). This shift can be considered either by the introduction of the corresponding phase shifts into the reference (modulating) voltages of low frequency, or by rotation through the appropriate angle of the scale of bearings.

The energy factor of ducts in the course of time is changed as a result of ageing, the effect of temperature, change in the load of the tubes and other reasons, which produces change phase displacement of both carrier and envelope.

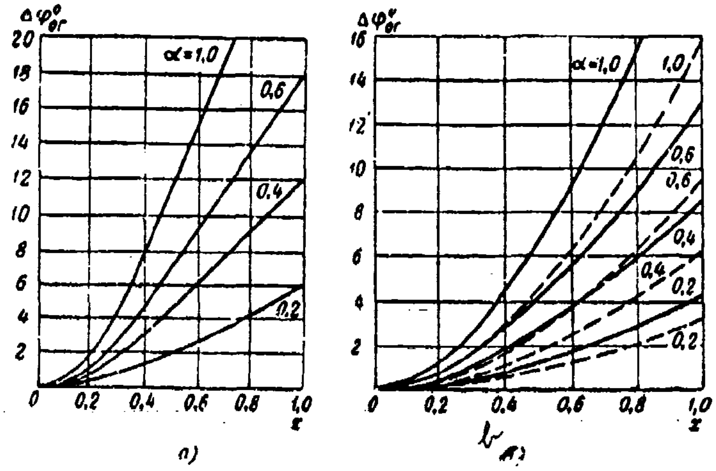

Fig. 8.32. Dependence of phase displacement of envelope on detuning for tuned amplifiers with single ducts with different $x = 2F/B$: a) for $n = 1$; b) for $n = 2$ (unbroken curves) and $n = 4$ (dashed curves).

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This variable part of phase displacement cannot be compensated for, if in circuit is not provided the possibility of frequent checking and adjustment of phase displacement.

During tuning of signal, any deviation from resonance is led to a change in the phase shift. Let us designate a change in phase displacement of envelope with detuning by $\Delta\varphi_{or}$; then

$$\Delta\varphi_{or} = \varphi_{or\ a=0} - \varphi_{or}(a).$$

Utilizing (8.36) and after designating $1+\eta^2=q$ and $2F/B = \alpha$, we will obtain

for a system from single ducts from (8.40)

$$\Delta\varphi_{or} = -\frac{n\alpha}{\psi(n)} \frac{x^2}{x^2+\psi^2(n)}, \qquad (8.42)$$

for a two-circuit circuit from (8.41)

$$\Delta\varphi_{or} = -\frac{2\alpha n}{\psi(n)}\left[\frac{x^4+x^2\psi^2(n)(4-3q)}{x^4q+2x^2\psi^2(n)\,q\,(2-q)+\psi^4(n)\,q^3}\right]. \quad (8.43)$$

The graph/diagrams of dependences $\Delta\varphi_{or}(x)$ for different resonance systems and dependences x(a) are given to Figs. 8.32-8.36.

The analysis of formulas (8.42) and (8.43) and curve/graphs shows that best for obtaining minimum $\Delta\varphi_{or}$ are the systems of single-circuit resonance and two-circuit band-pass amplifiers when $\eta<1$. Curve/graphs Fig. 8.36 make it possible to select the modulating frequencies.

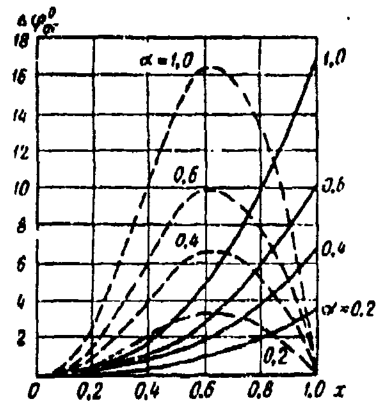

Fig. 8.33. Dependence of phase displacement of envelope on detuning for a single-stage band-pass amplifier when $\eta=0.5$ (continuous) and $\eta=1$ (broken) curved.

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During the rotation of the highly directional radiation pattern when the input of receiving indicator affects the momentum/impulse/pulse, duration and form of which depend on the rate of rotation and form of diagram, also can be observed the specific for this system errors of direction finding and distortion of the shape of pulse. As is known, the attack time at the output of tuned amplifier depends on waveform at input and on the transient characteristic of amplifier. If the attack time at input (the leading

impulse front) t_{nx}, and the transient response of entire circuit of receiving indicator is specific by the time of establishment t_y, then during satisfaction of the condition

$$t_{nx} > t_y \quad (8.44)$$

the form of output signal is close to that of the input signal. The resulting passband is selected on the basis of the condition

$$B \gg \frac{1}{t_y}. \quad (8.45)$$

The rise time of momentum/impulse/pulse can be approximately accepted equal to

$$t_{nx} = \frac{2\gamma_{0.1}}{2\pi F}, \quad (8.46)$$

where $2\gamma_{0.1}$ is width of radiation pattern at the level 0.1 from the maximum;

F - the frequency of the rotation of antenna.

Satisfying (8.44) and (8.45), they attain that, in order to the distortion of the form of the output pulse, the position of the axis of symmetry of which determines bearing, it did not lead to deterioration in the reading of bearing.

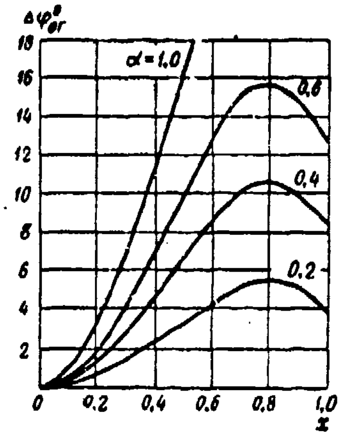

Fig. 8.34. Dependence of phase displacement of envelope on detuning for a band-pass amplifier with n = 2 and $\eta=1$.

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Analogous distortions occur during the rotation of cosinusoidal radiation pattern. They are developed in the fact that at output is obtained the more or less diffuse minimum, but not the zero values of the voltage. Besides possible wave form distortions during pulse advancing through the circuit of receiver display, will occur its time lag in time. During fine tuning for signal frequency, maximum and constant/invariable time lag can be taken into account by the introduction of the corresponding correction to bearing by the rotation of scale or deflection system of the cathode-ray tube of

indicator. However, time lag in the course of time changes, that it does not make it possible to consider it completely and a change in the envelope phase.

Imprecise tuning to signal will lead to a change in the time lag and respectively to bearing error. Time lag in amplifier is determined by the slope/transconductance of its phase response [2.7, 2.8]

$$t_{\text{з}} = \frac{d\varphi(\omega)}{d(\Delta f)}.$$

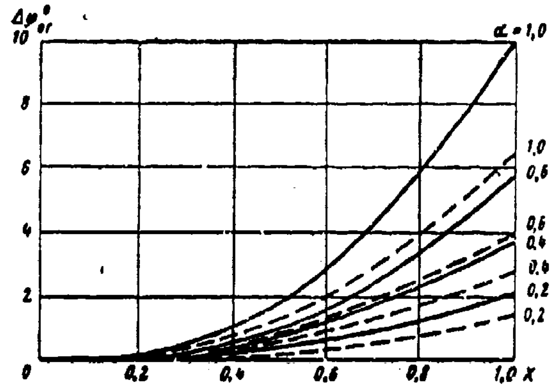

Fig. 8.35. Dependence of phase displacement of envelope on detuning for band-pass amplifier when $\eta=0.5$ and n = 2 (continuous) and when $\eta=0.5$; n = 4 (dotted curves).

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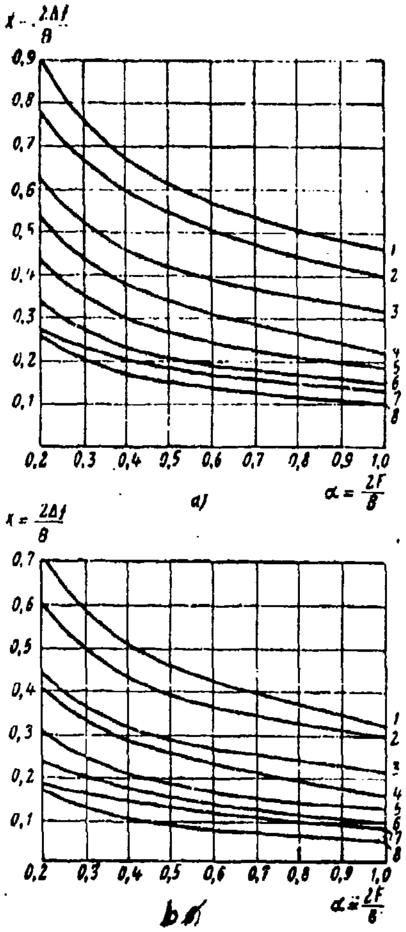

Fig. 8.36. Dependence of the permissible detuning $x = 2\Delta f/B$ on the ratio of the modulating frequency to the half of passband ($\alpha = 2F/B$) with the assigned errors of direction finding for the different resonance systems: a) the error of direction finding $\Delta = 1°$: 1 - band-pass amplifier, $\eta = 0.5$. $n = 4$; 2 - band-pass amplifier, $\eta = 0.5$. $n = 2$; 3 - band-pass amplifier, $\eta = 0.5$. $n = 1$; 4 - tuned amplifier, $n = 4$; 5

- tuned amplifier, $n = 2$; 6 - tuned amplifier, $n = 1$; 7 - band-pass amplifier, $\eta=1$, $n = 1$; 8 - band-pass amplifier, $\eta=1$, $n = 2$; b) the error of direction finding $\Delta = 0.5°$ (designation of curves the same as in 8.36a).

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A change of the time lag (Δt) with detuning Δf within the limits of passband B connected with phase displacement by enveloping dependence $\Delta t_1 = \frac{\Delta\varphi_{\text{гр}}}{2\pi F}$ can be determined:

for the single-stage tuned amplifier

$$\Delta t_1 = \frac{x^2}{1+x^2}\,\frac{1}{B\pi}, \qquad (8.47)$$

for an n-cascade tuned amplifier

$$\Delta t_n = \frac{n}{\psi(n)}\,\frac{x^2}{x^2+\psi^2(n)}\,\frac{1}{B\pi}, \qquad (8.48)$$

for an n- cascade band-pass amplifier

$$\Delta t_n = \frac{2n}{\psi(n)}\left[\frac{x^4 + x^2\psi^2(n)(4-3q)}{x^4 q + 2x^2\psi^2(n)\,q\,(2-q) + \psi^4(n)\,q^2}\right]\frac{1}{B\pi}. \qquad (8.49)$$

Minimum changes of the time lag with detunings within the limits of band B will be during the use of single-circuit resonance and two-circuit band-pass amplifiers when $\eta<1$.

In multistage amplifiers with B = const because of the expansion of the passband of each cascade/stage Δt_n it can be even less Δt_1. Knowing the frequency of the rotation of antenna F in hertzes and change of the time lag Δt in seconds, it is possible to easily determine the obtained in this case bearing error in the degrees:

$$\Delta=360F\Delta t.$$

For example, with detuning x = 0.5 and B = 2000 Hz for a single-stage tuned amplifier Δt_1 = 0.2 ms. If the frequency of the rotation of the radiation pattern of 20 Hz, this leads to bearing error $\Delta_t=1.44°$. Under the same conditions for a four-stage tuned amplifier Δt_4 = 0.075 ms and $\Delta_t=0.53°$.

Time lag and change in the time lag of the frequency modulated signal also can be found by formulas (8.46)-(8.49).

8.6. Phasometric radio direction finders.

In phasometric radio direction finders the bearing is determined from the envelope phase of the amplitude modulation which is obtained during the long running of radiation pattern (§ 2.3).

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Are known different methods of formation and rotation of the radiation pattern. Obtaining and rotation of cardioid or sinusoidal diagram can be realized with the aid of the rotatory antenna with sinusoidal diagram and phased motionless omnidirectional antenna [8.29], either by rotation of special reflector around motionless antenna [8.28], or by rotation of the rotor of goniometer with motionless antennas [8.14, 8.29].

The rotation of narrow-lobe diagram widely is applied in wide-base direction finders of short and ultra short waves and in UHF direction finders [8.25].

Is applied also the method of the rotation of radiation pattern with motionless antennas with the use of special electronic circuits.

Direction finders with the mechanical rotation of radiation pattern.

The rotation of radiation pattern causes modulation of the

oriented signal in amplitude with the frequency Ω, velocity-dependent of rotation and form of radiation pattern. The shape of the envelope of this modulated signal is determined by the form of radiation pattern, and phase - by direction in the oriented radio station (by angle of arrival of wave θ). The reference point of envelope phase is usually determined by the torque/moment of the passage of maximum or minimum of radiation pattern of the direction in north. If the strength of the field of the oriented signal $E = E_m \sin \omega t$, the antenna radiation pattern are determined by function $f(\theta)$ and the frequency of the rotation of radiation pattern is equal to Ω, then the voltage of signal on the input of receiving indicator in general form can be

$$u_c = U_m(\Omega, \theta) \sin \omega t. \qquad (8.50)$$

During the rotation of cardioid whose equation $f(\theta) = 1 - \cos \theta$, we obtain

$$u_c = U_m [1 - \cos(\Omega t - \theta)] \sin \omega t. \qquad (8.51)$$

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This modulated oscillation with modulation frequency Ω and modulation factor $M = 1$ (100o/o modulation), moreover phase displacement of envelope relative to reference point ($t = 0$) it is equal to the angle of arrival of wave θ (Fig. 8.37a). Such a modulation is observed only with ideal cardioid (§3.8), in real systems usually $M < 1$ and $u_c = U_m$

[1 - M cos (Ωt-θ) sin ωt.

If we consider that the receiving indicator is not introduced supplementary phase displacement of envelope and other errors (§8.5), then output potential of the linear amplitude detector of receiving indicator will be

$$u_{д} = U_{m\,д} + U_{m\,д} \cos(\Omega t - \theta).$$

Measuring with phasemeter value θ, we obtain the single-valued determination of bearing.

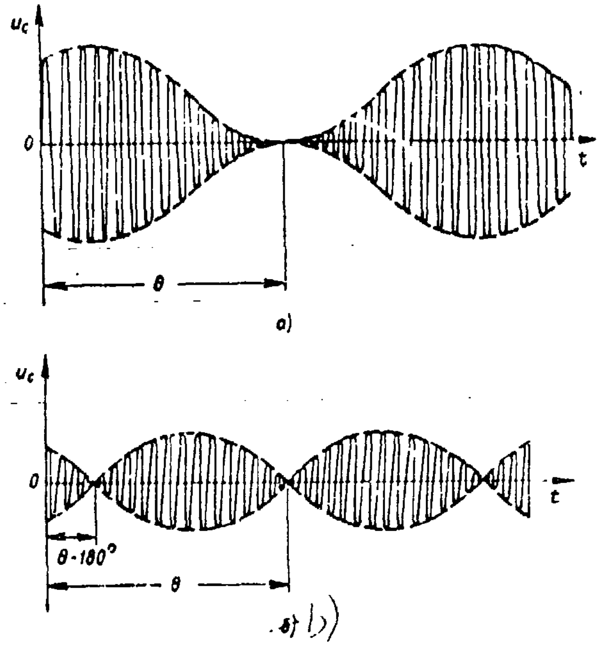

Fig. 8.37. Signal aspect with of the rotations: a) cardioid radiation pattern; b) sinusoidal radiation pattern.

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If the antenna radiation pattern is sinusoidal (eight), then the measurement of signal amplitude in time during the rotation of diagram will be it occurs according to the law

$$f(t, \theta) = \sin(\Omega t - \theta).$$

Signal at the input of receiving indicator takes the form

$$u_c = U_m \sin(\Omega t - \theta) \sin \omega t. \qquad (8.52)$$

These are modulated suppressed-carrier signal (Fig. 8.37b). during its detection the modulation frequency and phase displacement are doubled.

The reading of bearing on phasemeter is conducted with indeterminancy/uncertainty to 180°. For the elimination of ambiguity, is utilized the omnidirectional antenna.

The form of radiation pattern can differ significantly from sinusoid or cardioid. So, during the rotation of diagram with sufficiently narrow lug/lobe and distinct maximum modulation has pulse character (Fig. 8.38). In this case the bearing is determined by the phase angle between the reference point and the maximum value of momentum/impulse/pulse at the output of receiving indicator.

Usually on the axis of the rotation of antenna or goniometer, is located reference generator whose frequency is equal to the frequency of rotation, and phase is equal to zero during the passage of the minimum (or maximum) of radiation pattern of the direction in north.

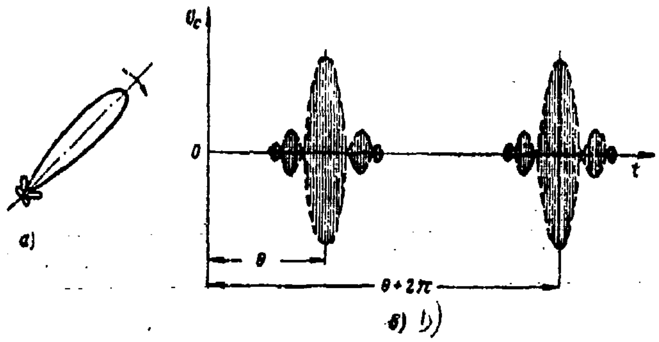

Fig. 8.38. Rotation of narrow-lobe radiation pattern: a) narrow-lobe diagram; b) signal aspect at the input of receiving indicator.

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Sometimes on the axis of rotation, is establish/installed selsyn transmitter, and receiving synchro serves for obtaining scanning/sweep on tube. Are applied sweep circuits with synchro-transformer and sine-cosine potentiometer [8.13]. Examples of the block diagrams of direction finders with the revolving radiation pattern are represented in Fig. to 8.39-8.41.

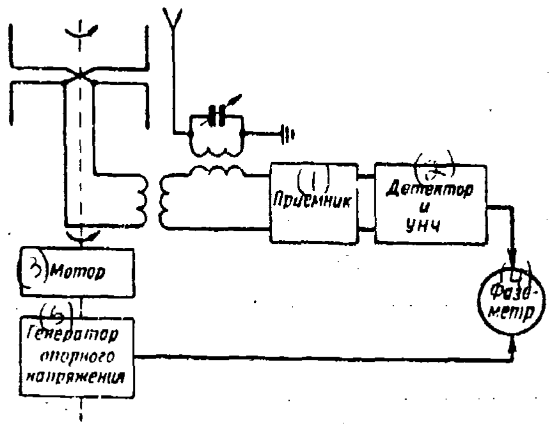

Fig. 8.39. Phase-meter radio direction finder with the rotatory antenna and arrow phasemeter.

Key: (1). Receiver. (2). Detector and UNCh [low-frequency amplifier]. (3). Motor. (4). Phasemeter. (5). Reference generator.

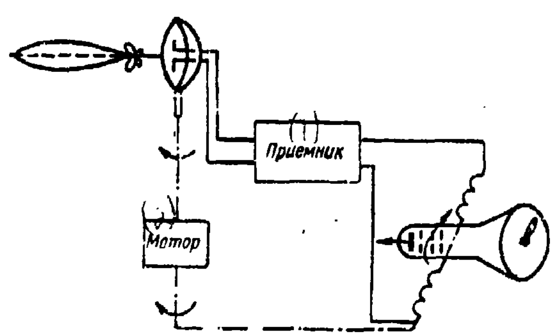

Fig. 8.40. Phase-meter radio direction finder with delineation of radiation pattern.

Key: (1). Receiver. (2). Motor.

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Cardioid pattern in direction finder (Fig. 8.39) is formed by the addition of the stress of the revolving H-type antenna and motionless omnidirectional antenna. the voltage of envelope, obtained as a result of detection, is fed to the arrow phasemeter to which will be feed/conducted also reference voltage from generator. The measured by phase meter phase difference of these stresses gives single-valued bearing in degrees.

Bearing is counted off with the aid of the phasemeters of different types. In this case, wide application found arrow phasemeters with balance detector and the instrument of direct current and phasemeters with cathode-ray tube.

Are applied also the circuits, in which is form/shaped the narrow pulse into the torque/moment of the passage of the sinusoid through zero in the direction of an increase in the current. The obtained momentum/impulse/pulse starts flip-flop. The analogous

momentum/impulse/pulse, obtained from reference voltage, inverts flip-flop.

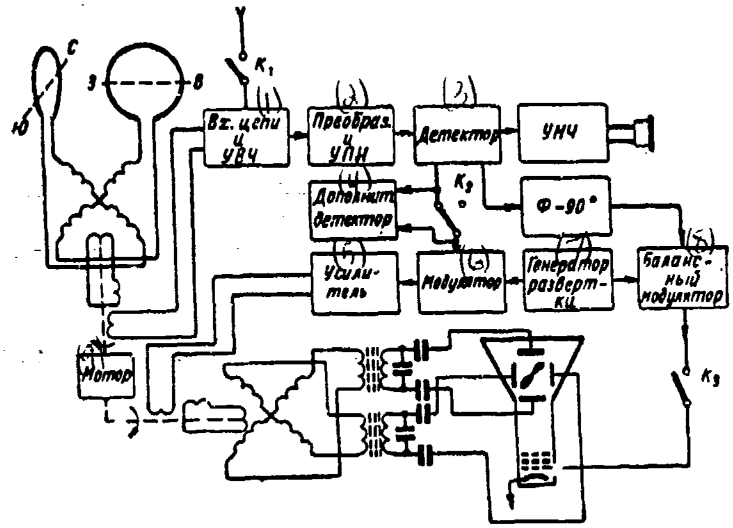

Fig. 8.41. Phase-meter radio direction finder with revolving goniometer.

Key: (1). Input of circuit and. (2). Converter and UPN. (3). Detector. (4). Will supplement. detector. (5). Amplifier. (6). Modulator. (7). Sweep oscillator. (8). Balance modulator.

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The time interval during which one of flip-flops tube is opened, and its average current are proportional to a phase difference in

measured and reference voltages [2.9]. The method of measurement of phase on the interval among the momentum/impulse/pulses, formed, as noted above, has a series of advantages in the relation to accuracy and possibility of the automation of the further treatment/working of bearings. It is described in §8.9.

In system (Fig. 8.40) rotates the antenna with acute/sharp radiation pattern. The output unidirectional voltage of receiver will be feed/conducted to the deflection coils of the cathode-ray tube which rotate synchronously and cophasally with antenna. With the reception of the signals of radio station on tube face, is drawn the antenna radiation pattern the axis of symmetry of which determines bearing.

Direction finder (Fig. 8.41) has the rotatable with the aid of high-frequency goniometer sinusoidal radiation pattern. The modulated with the frequency of rotation stress from search coil of goniometer is remove/taken with the aid of the revolving transformer. After conversion, amplification and detection in receiver, the obtained voltage of envelope in negative polarity is fed to the grid of the modulator where it modulates the sweep oscillator voltage (frequency its 15-80 kHz). The phase of the modulating stress depends on the azimuth of transmitter.

The voltage of sweep frequency enters through the revolving transformer the rotor of the goniometer of scanning/sweep which rotates in line with the rotor of high-frequency goniometer. The stator coils of goniometer of the scanning/sweeps in which in this case are formed two shifted to 90° voltages of sweep frequency, modulated with the frequency of the rotation of goniometer, are connected to the deflector plates of the tube on which is obtained the circular sweep.

The voltage in the stator coils of the goniometer of scanning/sweep is modulated by the voltage of the signal amplitude envelope whose phase depends on azimuth. Since the modulating voltage is utilized in negative polarity, on scope, is drawn reverse/inverse radiation pattern in the form of the propeller whose blade/vanes correspond to the minimums of figure-of-eight diagram.

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Along the axis of the symmetry of the obtained figure, is determined two-place bearing.

The side is determined by means of the closing/shorting of key/wrenches K_1, K_3 and of interrupting key/wrench K_2 (Fig. 8.41). During the closing/shorting of key/wrench K_1 to the input of

receiver, is fed the voltage of the omnidirectional antenna, matched on amplitude and phase with frame voltage. The obtained after detector voltage of the frequency of the rotation of goniometer during interrupting of key/wrench K_2 additionally is detected in negative polarity and modulates the sweep oscillator voltage.

Is simultaneous the voltage after the detector of receiver, out of phase to 90° with the aid of phase inverter (Φ - 90°), is fed to the input of the balanced modulator together with the sweep oscillator voltage. Output voltage from the balanced modulator during the closing/shorting of key/wrench K_3 is connected to grid - the modulator of the brightness of cathode-ray tube. On tube face, one of the lobes of the propeller of the image of bearing goes out, and can be counted off single-valued bearing.

Direction finders with the electrical rotation of radiation pattern.

The electrical rotation of radiation pattern usually is applied in systems with sinusoidal or cardioid radiation patterns (Fig. 3.42).

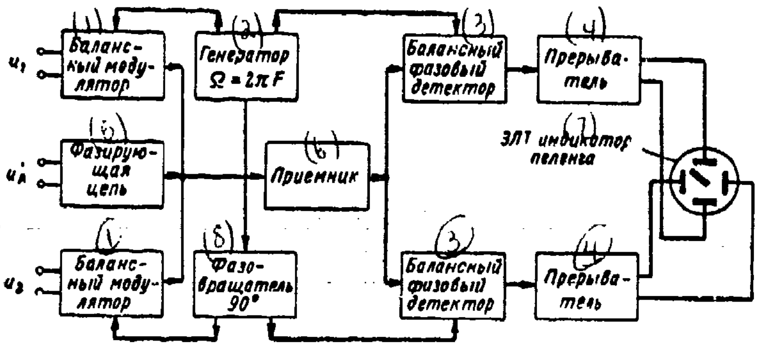

Fig. 8.42. Block diagram of radio direction finder with the electrical rotation of radiation pattern.

Key: (1). The balanced modulator. (2). Generator. (3). Balance phase discriminator. (4). interrupter. (5). Phasing circuit. (6). Receiver. (7). the indicator of bearing. (8). Phase inverter.

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For the rotation are utilized the balanced modulators to which are connected the voltages of the signal

$$u_1 = U_m \cos\theta \sin\omega t,\qquad (8.53)$$
$$u_2 = U_m \sin\theta \sin\omega t$$

and the modulating voltages

$$u_{\Omega 1} = U_{\Omega m} \sin\Omega t,$$
$$u_{\Omega 2} = U_{\Omega m} \cos\Omega t. \qquad (8.54)$$

Useful components at the output of the balanced modulators

$$u_{M_1} = U_M \cos\theta \sin\Omega t \sin\omega t, \qquad (8.55)$$
$$u_{M_2} = U_M \sin\theta \cos\Omega t \sin\omega t.$$

In the existing systems (Fig. 8.42) usually is applied the electrical rotation of cardioid pattern, for which is utilized also the voltage of the omnidirectional antenna, phased with the voltages of the directional antennas with the aid of the phasing circuit:

$$u_{\text{A}} = U_{m\,\text{A}} \sin\omega t.$$

During the addition of the output voltages of balanced modulators and voltage of the omnidirectional antenna on the input of receiving indicator, we obtain

$$u = U_{m\,\text{A}}[1 + M\sin(\Omega t + \theta)]\sin\omega t, \qquad (8.56)$$

where $M = \frac{U_M}{U_{m\,\text{A}}}$ is a coefficient of modulation.

The obtained expression characterizes the modulated in amplitude voltage, phase displacement of envelope of which relative to the phase modulating stress it is equal to bearing (θ). After conversion, amplification and detection, the signal of form $u_{\text{A}} = U_{m\,\text{A}} \sin(\Omega t + \theta)$ (is not considered as before phase displacement of envelope in receiver) is fed to balance phase discriminators [1.13], where as

supporting/reference will be feed/conducted voltages u_{g1} and u_{g2} (8.54).

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Through the load impedances balance phase discriminators, flow the direct currents, which create the voltages

$$u_{I}=U_{m\,c}\cos\theta;$$
$$u_{II}=U_{m\,c}\sin\theta.$$

Direct/constant voltages are utilized for an indication on cathode-ray tube just as in direction finder with audio modulation (§8.4).

In systems with the electrical rotation of radiation pattern, it is possible to apply large values Ω; the absence of the revolving cell/elements increases life and the reliability of system; direction finders with the electrical rotation of radiation pattern cheaper direction finders with mechanical rotation.

However, that that at entrance and exit of direction finder with electrical rotation are cell/elements, which require a precise balance (the balanced modulators and detectors), creates the supplementary sources of instrument errors because of the nonidentity of these cell/elements. Furthermore, the maintenance of the balance

of arms in the modulator circuits and detectors in the process of operation is very complex problem. In spite of these deficiency/lacks, the circuits of direction finders with the electrical rotation of radiation patterns find a use, and work proceeds on their perfection/improvement. Specific for phase-meter radio direction finders with the electrical rotation of radiation pattern is the use of the modulating voltages, shifted on Faye to 90°.

Let us determine the error of direction finding due to deviation from 90° phase shift of the modulating voltages.

During the deflection of the phase shift between voltages $u_{\Omega 1}$ and $u_{\Omega 2}$ from 90° on the angle φ of output potential of the balanced modulators they will take the form

$$u_{M_1} = U_M \cos\theta \sin\Omega t \sin\omega t,$$
$$u_{M_2} = U_M \sin\theta \cos(\Omega t + \varphi) \sin\omega t.$$

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Total voltage on the input of receiver in this case will be

$$u = U_{m\,a}[1 + MA\sin(\Omega t + \Phi)]\sin\omega t,$$

where

$$M = \frac{U_м}{U_{m\,\Omega}}, \quad A = \sqrt{1 - \sin 2\theta \sin \varphi};$$

$$\Phi = \operatorname{arc\,tg} \frac{\sin\theta\cos\varphi}{\cos\theta - \sin\theta\sin\varphi}. \tag{8.57}$$

The phase shift of envelope (Φ) is the function of both azimuth θ and angle ϕ.

Output potential of the detector of the receiver is equal

$$u_д = AU_{m\,д}\sin(\Omega t + \Phi).$$

If the reference voltages, supplied to the phase discriminators, have the same supplementary phase shift ϕ, i.e.,

$$u_{\Omega 1} = U_{m\,\Omega}\sin\Omega t \text{ и } u_{\Omega 2} = U_{m\,\Omega}\cos(\Omega t + \varphi),$$

[и = and]

that the output voltages of balance phase discriminators will be

$$u_1 = AU\sin\Phi,$$
$$u_2 = AU\cos(\Phi - \varphi).$$

The angle of deflection of the scanning/sweep of ray/beam on the tube of indicator will be defined as

$$\operatorname{tg}\alpha = \frac{\sin\Phi}{\cos(\Phi - \varphi)} = \operatorname{tg}\theta, \quad \alpha = \theta.$$

In this case, there is no error of direction finding. Depending on value ϕ = const with a change in the azimuth θ, is changed by A and respectively the amplitude of the beam deflection on tube, that it

does not lead to bearing error.

If in reference voltages there is no phase displacement φ, i.e.,

$$u_{\Omega 1}=U_{m\Omega}\sin\Omega t \text{ и } u_{\Omega 2}=U_{m\Omega}\cos\Omega t,$$

[и = and]

the output voltages of balance detectors are defined by the formulas

$$u_1=AU\sin\Phi \text{ и } u_2=AU\cos\Phi,$$

[и = and]

and the angle of deflection of ray/beam on tube will be defined as tg α = tg Φ, where α = θ + Δ; Δ is a bearing error.

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From formula for Φ (8.57) it is possible to find error Δ:

$$\operatorname{tg}\Delta=-\frac{\sin\varphi-\sin 2\theta+\sin(2\theta-\varphi)}{1+\cos 2\theta+\cos\varphi-\cos(2\theta-\varphi)}.$$

For small φ the error Δ is small and is determined by the formula

$$\Delta\approx\frac{\varphi(1-\cos 2\theta)}{2-\varphi\sin 2\theta}.$$

Maximum error will be at $\theta=90°$, 270°, it is equal to φ.

The fields of reradiation, which are located in quadrature with ground field, cause the errors in phase-meter radio direction finders

both with mechanical and with the electrical rotation of radiation pattern. The character of errors is completely analogous to the character of the errors, caused by the same reason in fixed loop radio compasses (§8.1). Their value can be estimated by formulas (8.8) and (8.9).

Selection of the frequency of rotation (modulation) in phase-meter direction finders.

From the viewpoint of obtaining minimum errors due to a change in phase displacement of envelopes during imprecise tuning and the achievement of high interference shielding from the signals of the mixing radio stations most desirable are small modulation frequencies.

However, in this case are limited the possibilities of direction finder on the stable direction finding of pulsed and short-term transmissions. For obtaining high-definition displacement in systems with the tracing of radiation pattern, the duration of the premise/impulse of telegraph signal must be 5-6 times more than cycle time of rotation. So, for the duration of normal telegraph premise/impulse 20 ms, the frequency of rotation must be order 250

Hz. The direction finding of short-term momentum/impulse/pulses with large porosity is possible, if they follow with high repetition frequency (several times higher than frequency of intermodulation).

During the rotation of narrow-lobe diagram, minimum duration of signal is equal to the period of turn.

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The frequency of modulation must be selected also from the conditions of the absence of the coincidence of the frequencies of intermodulation and modulation of signal. However, experiment shows that this criterion on is definite. The selection of the frequency of modulation in the spectrum of word was not virtually led to the noticeable errors of direction finding, since the coincidence of the frequencies of internal and external modulations was random and short-term.

In the existing systems with the mechanical rotation of radiation pattern, the rotational speed of design considerations does not exceed 3000 r/min (modulation frequency 50 Hz). Therefore their possibilities on the direction finding of pulse and short-term transmissions are very limited. In the direction finders of short waves with the electrical rotation of radiation pattern, are utilized

the modulation frequencies from 50 to 250 Hz. In VHF direction finders of modulation frequency, they lie/rest at the range higher than spectrum of word and they reach 5-6 kHz.

For the purpose of an increase in the sensitivity, and also the provision for direction finding of the pulse transmissions in radio direction finders with the rotation of cosinusoidal radiation pattern, are applied selective systems at the frequency of rotation and the different circuits of the accumulation of signals [8.14, 8.34-8.36].

8.7. Radio direction finders with large-base antenna system.

Amplitude direction-finding method.

In large-base radio direction finders greatest application/use has motionless circular antenna system with antenna commutator, which makes it possible to carry out direction finding from all azimuths (0-360°) (see § 3.11). As was shown in § 5.3 and § 6.5, in large-base antenna system for a decrease in the errors due to the influence of environment and propagation it is expedient to take the separation of

antennas as possible large. however, an excessive increase in the diameter of a circle of the arrangement/permutation of antennas encounters structural/design and operating difficulties. At the same time, if an increase in the diameter of a circle of system to 4-5 λ is led to an essential improvement in the characteristics, then a further increase in the diameter manifests itself much smaller.

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Direction finding can be realized visual either for audition on the maximum, the minimum, or according to comparison method (§2.3). For the execution of comparison method, antenna commutator must simultaneously create two equivalent, displaced to certain angle, the groups of antennas, for example along lines AA_1 and BB_1 in Fig. 3.40. Bearing corresponds to the coincidence of equisignal line with direction in radio station. Is feasible the phase-meter method of reading, for which is required the high-spin motion of antenna commutator (Fig. 3.38 § 8.6).

Without the rotation of antenna commutator, is realized automatic direction finding in certain sector $2\Delta\theta'_{макс}$, in this case they are utilized two subgroups of antennas and two-channel receiving indicator. Formula for calculation $2\Delta\theta'_{макс}$ (3.85) is given in § 3.11. By the rotation of antenna commutator the sector of

direction finding is moved within limits of 360°. The determination of bearing consists of the measurement of a phase difference of the stresses of the antennas of subgroups. The methods of measurement of a phase difference are presented in § 8.6 and in the second section of this paragraph.

In the antenna commutator of radio direction finder, the electrical delay-line length between the separate plates of rotor is designed, on the basis of geometry by the antenna of system for certain mean angle of the slope of a front of wave β_c. For other angles of the slope β, appears high-altitude error. Value β_c is selected such so that the high-altitude errors for angles $\beta_{макс}$ and $\beta_{мин}$ would be approximately identical. Let us calculate high-altitude error Δ_n for by the circular antenna of system with the simultaneous use of all antennas with direction finding from the minimum or from maximum. To the reading of bearing corresponds the position of the rotor of antenna commutator, when $\theta'=\Delta_n$ and the differential voltage of antennas (3.97) is equal to zero. For differential voltage it is possible to be faceted one first term of series (3.97)

$$U_\Delta = 4E_0 h_{c0} J_1\left(\frac{2\pi}{\lambda} bA\right) \sin\left(\gamma + \frac{\pi}{n} - \alpha_0\right) \operatorname{cosec}\frac{\pi}{n} = 0, \tag{8.58}$$

where by formula (3.88), set/assuming $\sin\theta' = \sin\Delta_n \approx \Delta_n$,

$$\gamma = \frac{\Delta_{s}}{1 - \frac{\cos\beta_0}{\cos\beta}}; \qquad (8.59)$$

α_0 - the angle of the axis of the symmetry of commutator with the nearest clockwise antenna (see § 3.11).

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Let us designate 2η the angular width of the plate of the rotor of antenna commutator. We assume that there are no errors due to the disturbance/breakdown of the electrical lengths of the line of time delays. For the directions of radio stations, which are characterized by condition $\frac{2\pi}{n} - \eta > \alpha_0 > \eta$ from (8.58) and (8.59), we will obtain

$$\Delta'_{s} = -\left(\frac{\pi}{n} - \alpha_0\right)\left(1 - \frac{\cos\beta_0}{\cos\beta}\right). \qquad (8.60)$$

During calculation Δ_s for the directions when $0 < \alpha_0 < \eta$ or $\frac{2\pi}{n} \geqslant \alpha_0 \geqslant \frac{2\pi}{n} - \eta$, one should consider that into differential voltage (8.58) instead of a difference in emf of the antennas of the n-th and n/2-nd, calculated by (3.96) when m = n by the formula:

$$U_{n\Delta} = 4E_0 h_{e0} J_1\left(\frac{2\pi}{\lambda} bA\right)\cos(\gamma - \alpha_0),$$

will enter voltage difference of identical phase (respectively α_0 = 0): each of which is proportional to the capacitance/capacities of the communication/connection of the plate of the stator of the n-th

(and the n/2)antenna with the plates of the rotor of the antenna commutator:

$$U'_{n\Delta}=4E_0h_{e0}J_1\left(\frac{2\pi}{\lambda}bA\right)\cos\gamma\left(\frac{\alpha_0+\eta}{2\eta}-\frac{\eta-\alpha_0}{2\eta}\right)=$$
$$=4E_0h_{e0}J_1\left(\frac{2\pi}{\lambda}bA\right)\cos\gamma\frac{\alpha_0}{\eta}.$$

High-altitude error is designed from the expression

$$U_\Delta-U_{n\Delta}+U'_{n\Delta}=0$$

or

$$\sin\left(\gamma+\frac{\pi}{n}-\alpha_0\right)\operatorname{cosec}\frac{\pi}{n}-\cos(\gamma-\alpha_0)+\frac{\alpha_0}{\eta}\cos\gamma=0. \qquad (8.61)$$

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Assuming that

$$\sin\frac{\pi}{n}\approx\frac{\pi}{n},\ \sin\left(\gamma+\frac{\pi}{n}-\alpha_0\right)\approx\gamma+\frac{\pi}{n}-\alpha_0,$$
$$\operatorname{cosec}\frac{\pi}{n}\approx\frac{n}{\pi},\ \cos\gamma\approx\cos(\gamma-\alpha_0)\approx 1,$$

and also by taking into account (8.59), we will obtain from (8.61) initial expression for the calculation of high-altitude error Δ''_s with $0\leqslant\alpha_0<\eta$ $2\pi n\geqslant\alpha_0\geqslant 2\pi/n-\eta$ (i.e. for α_0 within limits $\pm\eta$)

$$\gamma=\alpha_0\left(1-\frac{\pi}{\eta n}\right),$$

whence

$$\Delta''_{в} = \alpha_0\left(1 - \frac{\pi}{n}\frac{1}{\eta}\right)\left(1 - \frac{\cos\beta_c}{\cos\beta}\right). \quad (8.62)$$

the maximum value of high-altitude error is obtained from (8.60) and (8.62) when $\alpha_0 = \pm\eta$:

$$|\Delta_{в\,макс}| = \left|\frac{\pi}{n}\left(1 - \frac{\cos\beta_0}{\cos\beta}\right)\left(1 - \frac{n}{N}\right)\right|, \quad (8.63)$$

where n - a number of antennas; N is a number of plates of the rotor of antenna commutator.

Figures 8.43 depicts a change in the high-altitude error depending on angle α_0 for case of n = 18, N = 36, $\beta_c = 30°$. Curve 1 corresponds β = 60°, curve 2 is designed for β = 0.

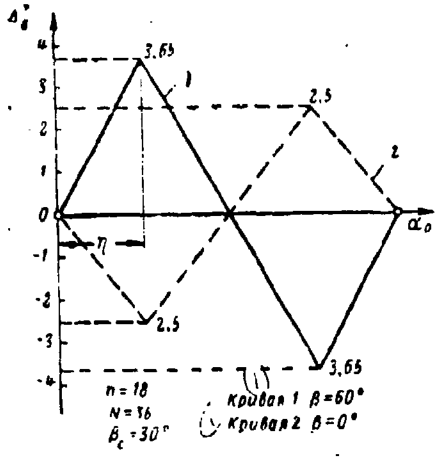

Fig. 8.43. High-altitude error of circular system.

Key: (1). Curved.

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With N = n, high-altitude error is not developed. For a decrease in the high-altitude error, it is expedient to decrease the number of plates of the rotor of antenna commutator. However, in order to achieve the smallest disturbance/breakdown of the designed electrical lengths of the lines of the time delays in the situations, when the plate of stator overlaps two plates of rotor, it is expedient to increase the number of plates of rotor. Therefore the selection of the number of plates of rotor is a compromise of the indicated opposite requirements.

So that the shunting of the lines of time delays with the loads of the feeders of antennas would not change calculated lengths, load impedance must be 4-5 times more than line characteristic [3.4].

In large-base radio direction finder besides by the circular antenna of system it is possible to use the rotatable rectilinear series of antennas with the distance between extreme antennas, long wave lengths. The method of the reading of bearing can be undertaken by any, used in rotatable antenna to system (visual or auditory reading on the minimum or on maximum during the manual rotation of

antennas, phase-meter during rotation by motor or other).

Including in the antennas of a rectilinear series the variable time delays, it is possible to carry out with motionless antennas rotation of the radiation pattern in certain sector. Finally, with established/installed motionlessly such an array of antennas, usually with reflector, is realized sector direction finding according to the principle of the measurement of a phase difference of the stresses of the halves of the antennas of system. The sector of direction finding in both cases is limited to the allowed values of high-altitude error and distortions of the form of radiation pattern.

Phase-difference direction-finding method.

Let us examine phase radio direction finders with motionless antennas. Phase-difference direction-finding method with four spaced antennas is described in § 2.3. In the case when it is required to determine only azimuth θ, are applied two spaced antennas.

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If 2b - the separation of two antennas, ϕ - phase displacement of their emf, θ - azimuth relative to perpendicular to the line of the arrangement/permutation of antennas, β - the angle of the slope of a front of wave, then

$$\varphi = \frac{4\pi}{\lambda} b \cos\beta \sin\theta$$

or

$$\theta = \arc\sin \frac{\varphi}{\frac{4\pi}{\lambda} b \cos\beta}$$

and

$$\beta = \arc\cos \frac{\varphi}{\frac{4\pi}{\lambda} b \sin\theta}.$$

The dependence θ and β on ϕ is nonlinear. Only at low values θ and $\beta \approx \beta_0$

$$\theta = \frac{\varphi}{k'_n}, \qquad (8.64)$$

where k'_n is a scaling factor from ϕ to θ:

$$k'_n = \frac{4\pi}{\lambda} b \cos\beta_0.$$

For obtaining high accuracy in phase radio direction finders, is taken the separation of antennas much larger than wavelength. Actually, if we designate: $\Delta\phi$ - the accuracy of reading of a phase difference of the stresses of antennas, $\Delta\theta$ - the accuracy of the

measurement of azimuth, $\Delta\beta$ - the accuracy of the measurement of high-altitude angle that,

$$\Delta\theta = \frac{\Delta\varphi}{\frac{4\pi}{\lambda} b \cos\theta \cos\beta}, \quad \Delta\beta = \frac{\Delta\varphi}{\frac{4\pi}{\lambda} b \sin\theta \sin\beta},$$

i.e. $\Delta\theta$ and $\Delta\beta$ they are improved with an increase $2b/\lambda$.

the best accuracy of the measurement of azimuth θ is obtained at $\theta = 0$ either 180°, the best accuracy of angle measurement of the slope of a front of wave β - when $\theta = 90$ or 270°. When $2b/\lambda > 1$ appears the multiformity in reading θ and β.

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For the resolution of multiformity, is applied one or several supplementary antenna systems with smaller separation with the phase-difference or other method of reading.

As it was mentioned, according to phase principle is realized sector direction finding in by the circular antenna to system and in the system of the antennas, arrange/located on straight line.

Besides the described in § 8.6 methods of measurement of a phase difference, are applied even compensation methods.

Figures 8.44a depicts the block diagram of the compensative method of measurement with manual control. In the channel of amplification of one of the voltages, is included the phase inverter. The circuit of comparison makes it possible to convert both compared voltages so as from indicator to find the position of the phase inverter when the phases of both stresses coincide. Sometimes in one of the channels for the selected method of measurement of phase coincidence, it is required to still provide the equalization of the amplitudes of stresses. As the circuit of comparison are applied the cascade/stage of sum or difference, the voltage converter into momentum/impulse/pulse, etc. As the indicator of phase coincidence can serve dial instrument or cathode-ray tube.

Figures 8.44b gives the block diagram of the automatic compensation for a phase difference with servo system. Phase inverter rotates by the motor which is supplied by the current of the circuit of control.

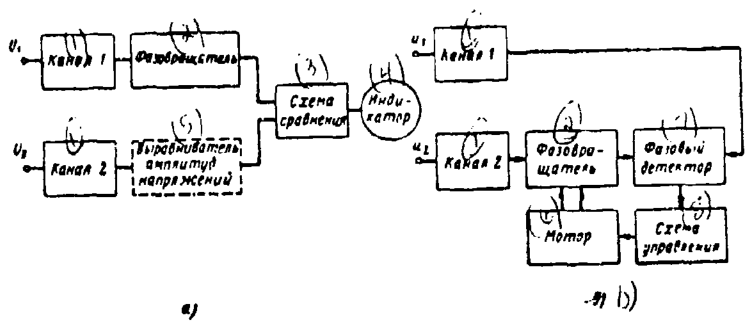

Fig. 8.44. Block diagram of the compensation method of the reading of the phase: a) with manual control; b) with servo system.

Key: (1). Channel. (2). Phase inverter. (3). Comparison circuit. (4). Indicator. (5). Equalizer of the amplitudes of stresses. (6). Motor. (7). The phase discriminator. (8). Circuit of control.

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The circuit of control develops voltage only if the current of the phase discriminator is not equal to zero.

As is known, the current of the phase discriminator is determined by product $I_{фд} = AU_{1вых}U_{2вых}\sin\varphi$ and $I_{фд} = 0$, when $\phi = 0$.

The automatic circuit of compensation possesses inertia and therefore it is applied when it is not possible to expect a very rapid change in the phases. The compensation methods of comparison of phases have accuracy of measurement on the order of 1°. It is possible to raise accuracy by means of frequency multiplication, but in this case measurement within limits of 360° becomes many-valued.

In the two-channel receiving indicators of phase radio direction finder, can be applied sum-differential method of the reading of bearing on cathode-ray tube.

The dissimilarity of the factors of amplification of the channels of receiver in amplitude and in phase after sum-difference block/module/unit is led to the errors of reading and to the ellipticity of image, examined into § 8.3. The azimuth error of phase radio direction finder k_n times is less than the error of reading according to the scale of the cathode-ray tube where

$$k_n = \frac{2\pi}{\lambda} b \cos\beta = 0{,}5k'_n.$$

Let us examine the effect of the nonidentity of the channels of receiver to sum-and-difference block/module/unit.

Let us designate: $E_0 h_e e^{-j\frac{2\pi}{\lambda} b_e \cos\beta \sin\theta}$ and $E_0 h_e e^{j\frac{2\pi}{\lambda} b_e \cos\beta \sin\theta}$ - voltage on the input of the channels of receiving indicator, b and $ce^{j\Phi}$ - voltage on the input of sum-and-difference block/module/unit after amplification in channels, $2b_e$ - the separation of the equivalent pair of antennas, E_0 - the strength of field at the center of system.

Then the sum of voltages on the input of sum-differential block/module/unit

$$U_\Sigma = b + ce^{j\Phi} = (b + c\cos\Phi) + jc\sin\Phi. \qquad (8.65)$$

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The difference in voltages, out of phase 90°, will be

$$U_\Delta = j(b - ce^{j\Phi}) = c\sin\Phi + j(b - c\cos\Phi). \qquad (8.66)$$

Is further included cathode-ray tube.

Equation for an image on cathode-ray tube face can be written in the form

$$U_\Sigma \cos\alpha + U_\Delta \sin\alpha = L$$

or

$$L = [(b + c\cos\Phi) + jc\sin\Phi]\cos\alpha +$$
$$+ [c\sin\Phi + j(b - c\cos\Phi)]\sin\alpha. \qquad (8.67)$$

Formula (8.67) is the equation of ellipse.

We will use expressions (III.3) and (III.6) of appendix III for determining the angle of the slope of the transverse $\alpha_{МНН}$ and of the relation of semiaxes of ellipse A/B.

On the basis of comparison of (III.1) and (8.67) we find

$$l = b + c\cos\Phi,\quad m = c\sin\Phi,\quad n = -c\sin\Phi,$$
$$p = -(b - c\cos\Phi).$$

From (III.3) we have

$$\operatorname{tg} 2\alpha_{МНН} = \frac{2(ln + mp)}{n^2 + p^2 - l^2 - m^2} = \operatorname{tg}\Phi. \qquad (8.68)$$

i.e. independent of the amplitude ratio of stresses on the input of the sum-and-difference block/module/unit

$$\alpha_{МНН} = 0{,}5\Phi. \qquad (8.68')$$

Thus, the angle of the orientation of the transverse of image on cathode-ray tube is always equal to one-half angle of a phase difference of the input voltage of sum-and-difference block/module/unit.

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The ratio of the semi-axes of ellipse from (III.6) is expressed by the formula

$$\frac{A}{B}=\frac{2(mn-l_p)}{l^2+m^2+n^2+p^2+\sqrt{(n^2+p^2-l^2-m^2)^2+4(ln+mp)^2}}=\frac{b-c}{b+c} \quad (8.69)$$

independent of phase displacement Φ.

Supplementary phase displacement of voltages on the input of sum and difference block/module/unit because of identity of channels

$$\Delta\Phi=\Phi-\frac{4\pi}{\lambda}b_3\cos\beta\sin\theta.$$

This phase displacement creates a change in the angle on the scale of the cathode-ray tube

$$\Delta\alpha=0,5\Delta\Phi=0,5\left(\Phi-\frac{4\pi}{\lambda}b_3\cos\beta\sin\theta\right) \quad (8.70)$$

and the azimuth error

$$\Delta\theta\approx\frac{\Delta\alpha}{k_n}=\frac{\Delta\alpha}{\frac{2\pi}{\lambda}b_3\cos\beta}.$$

Thus, bearing error is caused by the dissimilarity of the arguments (phases) of the amplification factors of the channels of receiver to sum and difference block/module/unit. The inequality of

the module/moduli of the factors of amplification of channels is led only to the ellipticity of image (8.69).

The reasons for phase displacement of voltages of channels can be different. In § 8.5 are given expressions for phase displacement of signal carrier frequency in single-channel receiver due to an inaccuracy in the tuning for signal frequency in the cases of the single and coupled circuits in the cascade/stages of amplification. If the channels of two-channel receiving indicator have identical selectivity curves and identical resonance frequencies, then an inaccuracy in the tuning for signal frequency will not lead to a supplementary phase difference of output voltages. Using the given in § 8.4 formulas, it is possible to calculate phase displacements in the different cases of the nonidentity of channels and during imprecise tuning for a two-channel receiving indicator.

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In [8.38], are carried out the calculations of a supplementary phase difference $\psi_{\text{макс}}$ of the output voltages of two-channel receiving indicator for the different cases of the nonidentities of channels and to imprecise tuning on the frequency of signal during application/use in the cascade/stages of single ducts.

In Fig. 8.45 are given the dependences of supplementary phase differences of output voltages $\psi_{\text{нако}}$ from the number of cascade/stages with detuning within the limits of passband for several special cases:

$$
\begin{array}{l}
1 \text{ для } \xi=1{,}0 \text{ и } 2\Delta\omega=0{,}1B_{\text{ср}}, \\
2 \text{ для } \xi=1{,}0 \text{ и } 2\Delta\omega=0{,}2B_{\text{ср}}, \\
3 \text{ для } \xi=1{,}0 \text{ и } 2\Delta\omega=0{,}3B_{\text{ср}}, \\
4 \text{ для } \xi=1{,}1 \text{ и } \Delta\omega=0, \\
5 \text{ для } \xi=1{,}2 \text{ и } \Delta\omega=0,
\end{array}
$$

Key: (1). for. (2). and.

where ω_0, ω_1, ω_2 are signal frequencies and resonances of first and second channels, B_1 and B_2 - the bandwidth at the level 0.707,

$$\frac{B_1+B_2}{2}=B_{\text{ср}},\ \frac{B_2}{B_1}=\xi,\ \Delta\omega=\omega_2-\omega_1.$$

From curves it follows that most strongly is developed the dissimilarity of the resonance frequencies of the ducts of channels ($\Delta\omega$), the dissimilarity of passbands in channels (ξ) it manifests itself less.

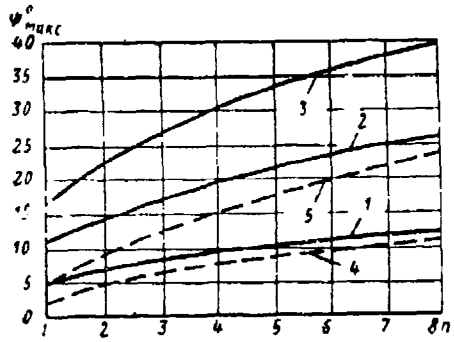

Fig. 8.45. Dependence of maximum phase differences on number of cascade/stages.

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Due to the nonidentities of channels and inaccuracy in the tuning, can appear large errors. Therefore in phase radio direction finder for obtaining high accuracy, it is necessary to systematically control and respectively to regulate the identity of the parameters of channels and their amplification factors in phase.

In the receiving indicators of phase radio direction finders, are applied the different methods of the complete or partial association of channels, based on the same principles, as in amplitude radio direction finders (see § 8.4).

Figures 8.46 depicts set-up with conversion of frequencies in channels into close in the rating of frequency. Let the voltages on the input of high-frequency amplifiers (UVCh) will be $U_1 \sin \omega t$ and $U_2 \sin (\omega\tau + \phi)$. To first mixer (SM1) of one of the channels, is given the voltage of the generator $U_r \sin (\omega_r t + \varphi_r)$. To the first mixer of second channel, is given the voltage from the output of the second mixer (SM2), obtained by the addition of the voltages of frequencies ω_r and Ω, moreover $\Omega \ll \omega_r$:

$$U''_r \sin [(\omega_r + \Omega) t + \varphi_r + \varphi_\Omega].$$

The voltages of close in rating intermediate frequencies $\omega'_\pi = \omega_r - \omega$ and $\omega''_\pi = \omega_r - \omega + \Omega$ are amplified in common/general/total amplifier (UPCh). After detector-mixer is obtained the voltage of frequency Ω. If in channels UVCh and in channels UPCh phase displacements of the amplified frequencies are identical, then on the output of detector-mixer voltage takes the form $U_s \sin (\Omega t + \varphi_r - \varphi)$. The measured phase difference in frequency Ω is equal to a phase difference ϕ in high frequency in the input of channels.

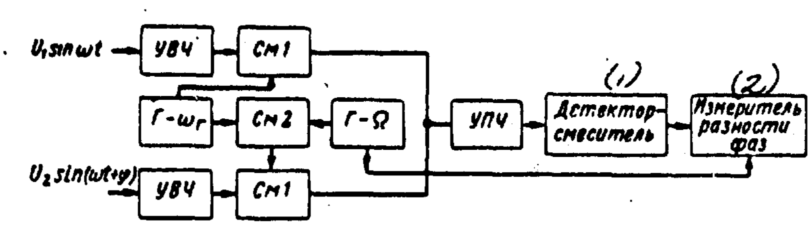

Fig. 8.46. Block diagram of two-channel receiving indicator with frequency conversion in channels into close in rating of value.

Key: (1). detector-mixer. (2). Meter of phase difference.

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In [1.14], is described the based on this principle more compound circuit of phase radio direction finder for the direction finding of artificial Earth satellites.

Figures 8.47 depicts the block diagram of single-channel amplification with the transformation of signals according to the method of heterodyning. One of the voltages of signal $U_1 \sin \omega t$ is fed to the inputs of two balanced modulators B1 and B2, to which will be feed/conducted also the voltage of the reference oscillation of

low frequency $U_\Omega \sin\Omega t$.

Voltage on input B1, where they enter voltage U_1 directly and voltage U_Ω with phase displacement 90° (in Fig. 8.47 cell/element Φ_2 - 90°), it will be

$$u_{\rm I} = U_1 \sin\omega t + U_\Omega \cos(\Omega t + \varphi_r).$$

Voltage on input B2, where enter voltage after phase displacement to 90° (Φ_1 - 90°) and voltage U_Ω directly, has the expression

$$u_{\rm II} = U_1 \cos\omega t + U_\Omega \sin(\Omega t + \varphi_r),$$

where φ_r is an initial phase of the voltage of reference frequency.

Output potential of the balanced modulators, connected to the input of receiver, will be

$$u_0 = kU_1U_\Omega[\sin\omega t\cos(\Omega t + \varphi_r) + \cos\omega t\sin(\Omega t + \varphi_r)] = \\ = U\sin[(\omega + \Omega)t + \varphi_r].$$

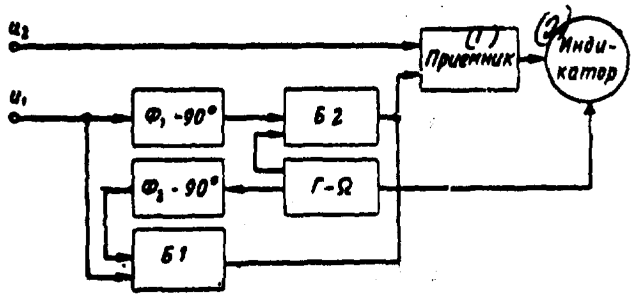

Fig. 8.47. Block diagram of single-channel receiving indicator with the transformation of signals according to the method of heterodyning.

Key: (1). Receiver. (2). Indicator.

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The second stress $U_2 \sin(\omega t + \phi)$ is fed to the input of receiver directly.

After square-law detection of voltages u_0 and u_2, is obtained the stress component of frequency Ω:

$$u_{\Omega\,\text{вых}} = U_{\text{вых}} \cos(\Omega t + \varphi_{\text{г}} + \varphi). \qquad (8.71)$$

A phase difference of stresses (8.71) and of reference oscillation is equal to a phase difference ϕ of the voltages of signal.

For the exception/elimination of crosstalk before the balanced modulators in channels, can be provided the high-frequency amplifiers. However, this complicates diagram. To diagram are characteristic the deficiency/lacks, connected with the application/use of nonlinear cell/elements in receiving circuit.

If in two-channel receiving indicator of phase radio direction finder, is used sum-and-difference block/module/unit, then the part of the amplification it is possible to realize before sum and difference block/module/unit according to phase principle, the part of the amplification after sum-and-difference block/module/unit - according to amplitude principle. In this phase-amplitude receiving indicator (see § 2.3), are somewhat lowered the requirements for the identity of separate cascade/stages and is facilitated the development of radio direction finder.

Let us examine an example. Let the amplification and selectivity

in receiving indicator be realized in essence six by cascade/stages of IF amplifier. We are assigned $k'_u=8$ and by permissible instrument error 0.2°. In Table 8.3 are designed the permissible nonidentities of the factor of amplification of one cascade/stage in amplitude and in the phase on the assumption that:

a) all the amplification carried out according to phase-difference method;

b) all the amplification carried out according to amplitude method;

c) are 2 cascade/stages according to phase method and 4 on amplitude.

As can be seen from Table 8.3, requirement for the identity of channels in the third version of receiving indicator somewhat they descend in comparison with the first two versions.

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In [8.24], are described the results of the tests of phase radio direction finder for frequency band 3-30 MHz with antenna system of two mutually perpendicular pairs of the vertical wire antennas (Fig.

2.11). For direction finding were utilized pairs 1-3 and 2-4 with separation 120 m and pairs 1-2 and 2-3 with separation 84.6 m with two-channel receiving indicator. For the resolution of multiformity, was applied goniometric radio direction finder with diverse vertical antennas. Were compared the results of direction finding on phase and goniometric systems.

The average quadratic instrument error of systems, determined by direction finding at frequencies 5-7 MHz of transmitter at a distance of approximately 3-5 km (terrestrial wave), render/showed:

for phase radio direction finder 0.2°,

for goniometric radio direction finder 1.5°.

With the direction finding of 95 remote radio stations in the range of frequencies 5-25 MHz, the readings were taken during 30-50 s for each performance. Mean square errors render/showed:

of phase radio direction finder with separation 120 m 1.24°, with separation 84.6 m 1.4°;

of goniometric radio direction finder 3.48°.

Is known the application/use of a phase radio direction finder in ultrashort-wave range (200 MHz) for radio-astronomical observations and the direction finding of the artificial Earth satellites and the spacecraft. In radio direction finder were utilized two parabolic antennas size/dimensions 8 x 18 and 11 x 22 m, spread up to distance on the order of 100 wavelengths when the lug/lobe of radiation pattern was equal approximately to 30 min. the accuracy of direction finding was ±1 min [8.11].

Table 8.3. Requirements for the cascade/stage of IF amplifier.

(1) Метод пеленгования	(2) Допустимая неодинаковость усиления	
	(3) по фазе, град	(4) по амплитуде, %
(5) Фазовый	0,6	(8) Любая
(6) Амплитудный	(8) Любая	±0,12
(7) Фазово-амплитудный	1,8	±0,18

Key: (1). Direction-finding method. (2). Permissible dissimilarity of amplification. (3). on phase, deg. (4). in amplitude, o/o. (5). Phase. (6). Amplitude. (7). phase-amplitude. (8). Any.

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Pulse radio direction finder.

With the direction finding of pulse transmissions, it is possible to measure not a phase difference of the stresses of antennas, but difference in time began the inductions of

momentum/impulse/pulses in diverse antennas of radio direction finder.

The simplest diagram of pulse radio direction finder with two diverse antennas is depicted on Fig. 8.48.

The voltages of antennas A_1 and A_2 with the aid of synchronously switching circuits PP1 and PP2 are fed to the input of receptor and from its output to vertical plates of cathode-ray tube. Voltage A_1 falls on the plates of tube directly, voltage A_2 - after calibrated delay unit and reverser of phase to 180°. Timer synchronizes the frequencies of switchings (PP1) and (PP2) and the frequency of sweep circuit which is connected to the horizontal plates of cathode-ray tube.

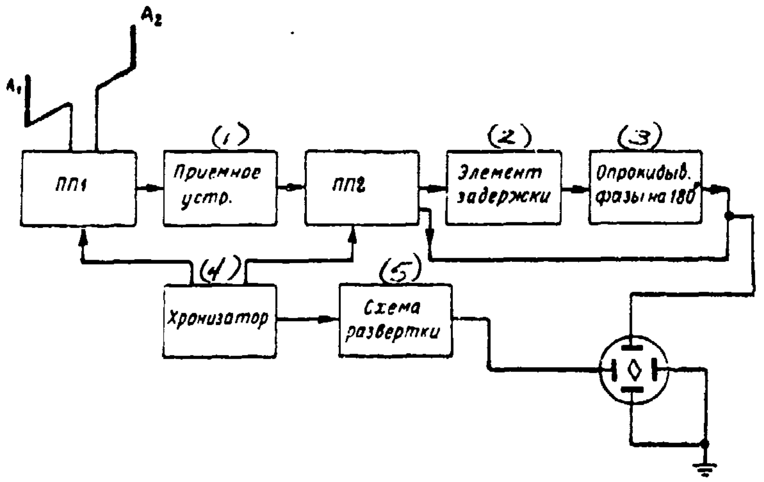

Fig. 8.48. Pulse radio direction finder.

Key: (1). Receiving arrangement. (2). Delay unit. (3). Reverser of phase to 180°. (4). Timer. (5). Sweep circuit.

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By delay unit they attain pulse coincidence. Time difference of the arrival of momentum/impulse/pulses at two antennas is counted off on delay unit, i.e.,

$$\Delta t = \frac{2b}{3\cdot 10^{8}} \sin \theta' \cos \beta.$$

Delay time determines bearing. Countdown is facilitated during the use of gauging time/temporary marks.

The accuracy of the coincidence of momentum/impulse/pulses can be sufficiently large, approximately 0.03 pulse durations. At the same time, the accuracy of direction finding is obtained low. In the directions where the accuracy of direction finding is greatest (about perpendicular to base), with error in time δt angular bearing error will be

$$\Delta = \frac{\delta t}{2b} 3\cdot 10^{8};$$

for example if δt = 0.03 μs, 2b = 500 m, $\Delta \sim 1°$.

This method is applied for the direction finding of atomospheric discharges at frequencies on the order of 10 kHz. In [8.25], is described the system of three spread up to distance 1/3 - 1/10 wavelengths of the vertical wire antennas 38 m in height. The determination of bearing is reduced to the measurement of time difference of the appearance of momentum/impulse/pulses of

atomospheric discaarges at the output of three identical amplifiers, connected to antennas. This system is named E - ϕ by system. The accuracy of direction finding is evaluated approximately at 0.5-1°.

8.8. Radio direction finders with cyclic measurement of phase in high frequency.

The operating principle of phase radio direction finier with the rotatory antenna is examined in chapter 2. The practical implementation of the long running of antenna causes considerable difficulties, since radius of gyration must be great in order to decrease the interference errors and the errors of local environment, but he rotational speed by sufficiently high, that would become possible the rapid direction finding.

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For overcoming the indicated difficulties in radio direction finders, continuous rotation of one antenna is replaced by the series connection of a series of the antennas, arrange/located in circumference (Fig. 8.49). It is posssible to change over antennas with the aid of mechanical or electron commutator.

On Fig. 8.50 is given the circuit of capacitive commutator. The stator of commutator is a series of plates 1, 2, ..., n, arrange/located in the circumference to which are connected the antennas. The plate of rotor is connected with receiver. It is possible to supply several receivers from one by antenna of system and to realize simultaneous direction finding of several stations, for which are provided several plates on rotor. Rotor rotates with the aid of motor. When the plate of rotor is located above any (for example, the k-th) plate of stator, is obtained the greatest capacitive coupling of receiver with this antenna. The phase of the stress, removed to receiver at this moment, corresponds to the phase of emf induced in the k antenna. When rotor turns itself and its plate will stop above the plate of following antenna (k + 1), removable on receiver voltage has the same phases, as emf (k + 1) antennas. In the intermediate positions of rotor, the receiver is connected with both antennas and voltage at the input of receiver gradually changes the phase from value, corresponding to emf in the k antenna, to value, corresponding to emf in (k + 1) antenna.

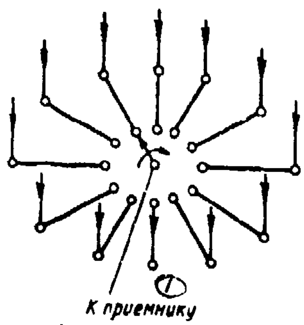

Fig. 8.49.

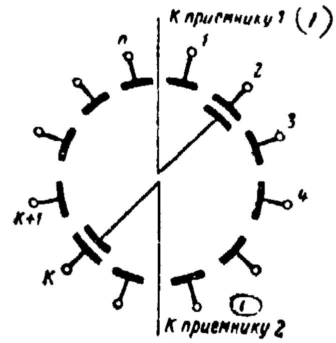

Fig. 8.50

Fig. 8.49. Switching circuit of antennas.

Key: (1). To receiver.

Fig. 8.50. Schematic diagram of capacitive commutator.

Key: (1). To receiver.

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Change of the phase of voltage on the input of receiver which would be obtained during the rotation of single antenna, is shown to dotted curve in Fig. 8.51. A change in the phase during the rotation of the rotor of commutator is represented in this same the figure of unbroken curve. This is the same sinusoid to which are superimposed the small oscillations, connected with the transition of the plates of rotor from communication/connection with one antenna to the next.

Electronic switching can be carried out by different means. Is most expedient the application/use of semiconductor diodes, which possess a small capacitance/capacity, small resistance to the current of forward direction and large resistance to countercurrent. An example of the circuit of electronic switching is represented in Fig. 8.52. Fach of the antennas is included to the input of the receiver through the same commutating circuit which is depicted in figure only for one antenna. Point A of the commutating circuit through resistor/resistance R_1 is connected with the impulser from which during entire period of T of commutation, with the exception only of short interval/gap τ, is fed negative voltage. Positive pulses with duration τ are fed consecutively to each antenna and during commutating period fall on whole n antennas.

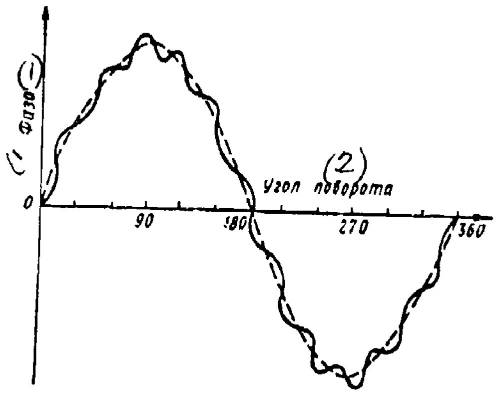

Fig. 8.51. Dependence of phase of output voltage from position of rotor of commutator.

Key: (1). Phase. (2). Angle of rotation.

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Negative voltage at point A cuts off diodes D_2 and D_1, cutting off the antenna circuit from the input of receiver and turning on in the

circuit of the antenna load resister R_2 and triggers the diode D_3, which closes point A to the earth. Throttle/choke L serves for the transmission of the direct current of diodes. Positive pulse makes diodes D_1 and D_2 those who carry out. Antenna is connected with receiver during short-circuited resistor/resistance R_2. Simultaneously is cut off diode D_3 and is removed short circuit to the earth. A change in the phase of stress on the input of receiver (Ris. 8.53) occurs irregularly in accordance with the connection of the new antenna through time intervals τ.

With any method of commutation in the input of receiver enters the voltage of the high frequency of alternating/variable phase, i.e., phase modulated. The period of modulation is equal to commutating period a the initial phase of modulation curve it is equal to bearing. Phase-modulated oscillation is also frequency modulated, since frequency, equal of derived phase on time, in alternating/variable phase will be variable.

For the manifestation of the phase of the curve of modulation, i.e., bearing, it is necessary to produce demodulation, isolating modulation curve.

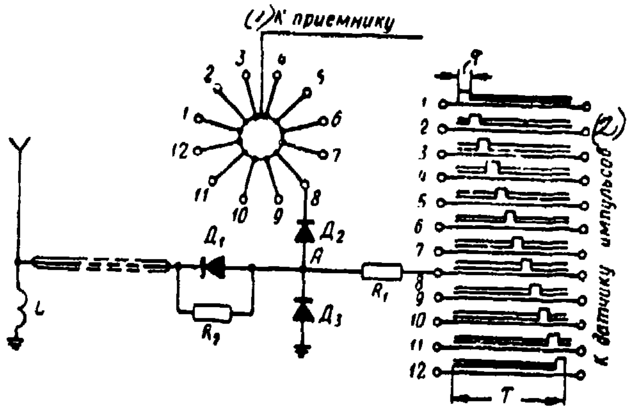

Fig. 8.52. Circuit of electron commutator.

Key: (1). To receiver. (2). To impulser.

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It is possible to produce demodulation in frequency, utilizing the FM discriminator, or on phase, utilizing the phase discriminator. Accordingly, are distinguished two basic versions circuits of the

radio direction finder in question.

Frequency change, caused by the displacement of observation point at constant velocity, he is called the Doppler effect. In the radio direction finders in question frequency change is obtained also as a result of the real or equivalent displacement of antenna. Therefore radio direction finders of the type in question, in which is utilized frequency modulation for the indication of bearing, call quasi-Doppler radio direction finders. Of two the methods described above of commutation, mechanical method is more adapted for use during frequency detection. Actually, Fig. 8.51 shows that the derivative of phase in terms of time, i.e., frequency, changes in essence according to cosinusoidal law with the period, equal to the period of the rotation of commutator. On this fundamental curve are superimposed the frequency variations with the period, equal to transit time from one antenna to following. These oscillations subsequently easily can be filtered out. In the case of electronic switching (Fig. 8.53) the derivative is turned theoretically into infinity at the torque/moments of switching antennas and it is equal to zero (i.e. frequency is constant) in the interval/gaps between switchings.

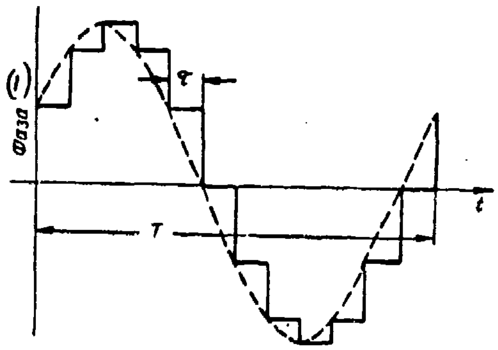

Fig. 8.53. Change in phase of output voltage with electronic switching.

Key: (1). Phase.

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Virtually transient phenomena during switching occur not instantly and the overshoots of frequency have the finite quantity and the final duration, which depend on the character of transient process. The use of the obtained frequency modulation causes considerable

difficulties.

The realization of phase demodulation is possible only by the path of the comparison of two oscillations a phase difference of which is isolated by the phase discriminator.

During the description of the operating principle of radio direction finders with the rotatory antenna and the phase-difference method of reading (§ 2.3) was examined the easiest method - comparison of the phase of the stress of the rotatory antenna with the phase of the stresses of motionless antenna. Usually is applied another method: are compared the phases of the stresses in two adjacent antennas. For this purpose, the voltage, removed from any antenna (for example, the k-th), is delayed by the filter of time delay of time τ. This delayed voltage and the undelayed voltage (k + 1)-1 antennas are compared between themselves on the phase discriminator. The radio direction finders in which is utilized phase modulation by means of the comparison of the voltages of two adjacent antennas, they are called differential-phase radio direction finders. Electronic switching makes it possible to carry out a differential-phase direction finding simpler than mechanical.

Let us examine the considerations, which determine the selection of the main parameters of phase radio direction finder with the

switched antennas.

Diameter of a circle along which are arrange/located the antennas, it is desirable to make largest possible for a decrease in the local and interference errors. In § 5.3 it is shown, that an increase in the separation of antennas up to $2R/\lambda = 3-4$ sharply decreases local errors. A further increase in the separation smaller affects local errors.

Final dimension is establish/installed as compromise between the given requirement and structural/design and operating requirements.

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For determining the necessary number of antennas, let us examine the spectrum of signal during the rotation of single antenna. The voltage, removed to the input of receiver, is equal (2.19) with $\beta = 0$

$$e = E_m \sin[\omega t + mR\cos(\Omega t - \theta)]. \quad (8.72)$$

From the theory of frequency modulation, it is known that the spectrum of the frequency-modulated or phase signal consists of the infinite series of the side frequencies, which differ in frequency from the carrier on $\pm k\Omega$, where k is a number of harmonic, and that the essential for reproduction side frequencies range from 0 to $M\Omega$,

where M is an index of modulation. In this case $M=mR$. Utilizing the changed over antennas, we replace continuous modulation curve together of its discrete values at the points of antenna location. According to Kotelnikov theorem for the reproduction of curve to the limited spectrum together of its discrete values the distance between discrete points Δt must satisfy the condition

$$\Delta t<\frac{1}{2f_c}, \qquad (8.73)$$

where f_c is a boundary of the spectrum of curve.

Assume $f_c=\frac{M\Omega}{2\pi}$, we obtain for angular distance ϑ_0 between the antennas

$$\vartheta_0=\Omega\Delta t<\frac{\Omega}{2f_c}=\frac{\pi}{M}=\frac{\lambda}{2R}$$

and for the linear distance d

$$d\approx R\vartheta_0\leqslant\frac{\lambda}{2}. \qquad (8.74)$$

Thus, the distance between antennas must be less than the half of wavelength. In the case of differential-phase direction-finding method, this condition provides the absence of multiformity. Actually, the difference of phases ψ of the voltages of two antennas, arrange/located at angles ϑ_k and ϑ_{k+1}, it is equal to

$$\psi=mR\sin(\vartheta_{k+1}-\theta)-mR\sin(\vartheta_k-\theta)=$$
$$=2mR\cos\left(\frac{\vartheta_{k+1}+\vartheta_k}{2}-\theta\right)\sin\frac{\vartheta_{k+1}-\vartheta_k}{2}.$$

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The distance between adjacent antennas is equal

$$d=2R\sin\frac{\theta_0}{2}=2R\sin\frac{\theta_{k+1}-\theta_k}{2}.$$

Thus, we obtain

$$\psi=md\cos\left(\frac{\theta_{k+1}+\theta_k}{2}-\theta\right). \qquad (8.75)$$

For providing the uniqueness, a maximum phase difference must be less π and, therefore, $d < \lambda/2$. Virtually the distance between antennas takes order $\lambda/3$. It is possible to increase the distance between antennas to $d = \lambda$. This corresponds to the contraction of the band of the reproducible frequencies 2 times, which still barely affects the distortion of the form of curved phase modulation. For an exception/elimination in this case of the multivalence of readings is conducted the twofold differential comparison of the phases: voltage with differential phase (8.75) is delayed in the second filter of time delay of time τ and is compared with undelayed voltage [8.19].

The velocity of rotation or commutation of antennas is selected, proceding from the considerations, given in § 8.6.

As single antennas are applied symmetrical or asymmetric vibrators. The interconnections of working antenna with inoperative

cause supplementary changes in the phase of the stress in antenna. The frequency of these changes is higher than the frequency of fundamental phase modulation, and during sufficient filtration they do not affect bearing. Requirements for the degree of filtration are reduced with a decrease in the effect of the inoperative antennas. For this purpose, in the circuit of the inoperative antennas it is possible to involve relatively high resistor/resistances (see Fig. 8.52).

The passband of the frequencies of the receiver in high frequency must be selected in accordance with the width of the spectrum of signal, i.e., must be more than $2M\frac{\Omega}{2\pi}$. In quasi-Doppler radio direction finders must be provided sufficiently small phase displacement of modulating curve.

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The necessary passband is designed from data of § 8.5. In differential-phase radio direction finders phase displacement, common/general/total for both compared voltages, are not caused the errors.

Frequency stability of signal is especially important for differential-phase radio direction finders. During the divergence of

signal frequency from the nominal frequency of the filter of time delay, the signal carrier frequency undergoes in filter phase displacement. The value of phase displacement must not be great and in sum with the maximum value of alternating/variable phase must not exceed π. They usually limit phase displacement by carrying value 10-20°.

For operational provisions of the filter of time delay at stable frequency, it is possible to utilize the circuit of conversion of frequencies, in which the voltage frequency, which enters the filter of time delay, is equal to the frequency of the local oscillator. The latter is stabilized by quartz (see Fig. 8.55).

Passband from low frequency is determined basically by the time constant of indicator. It can be undertaken in accordance with the necessary velocity of search and trackings (§ 2.6).

The measurement of phase in output is conducted by one of the methods, described in § 8.6.

Because of the possibility of using the different switching circuits, the different methods of the measurement of phase and different structure of receiving indicator, is obtained the large number of versions of the fulfillment of radio direction finders with the cyclic measurement of phase in high frequency.

Figs. 8.54 and 8.55 show two variations of block diagrams [8.20, 8.21, 8.19]

Figures 8.54 depicts the block diagram of quasi-Doppler radio direction finder. motor revolves capacitive the commutator of antennas. On one shaft with motor, is located also generator of reference voltages, which creates two voltages of frequency Ω, shifted between themselves on phase to 90°. Initial phase of one of the reference voltages is such that voltage is passed through zero at that torque/moment when commutator realizes a maximum antenna coupling, which is found on the initial line of the calculation of angles (line north - south). The voltage of high frequency from commutator will be feed/conducted to amplifier (UVCh).

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Further it is converted in mixer (Sm) with the aid of the local oscillator and is amplified in amplifier of intermediate frequency (UPCh). The intensive voltage is limited in amplitude limiter and is detected by the FM discriminator. After amplification on low frequency, output voltage is compared with respect to phase with reference voltages. Bearing is read from indicator.

Figure 8.55 depicts the circuit of differential-phase radio direction finder. The electron commutator is controlled by impulser. Momentum/impulse/pulses from sensor come also electronic reference

generator, converting into them two sine voltages of the frequency of commutation, shifted one relative to another on phase to 90°. These two voltages supply indicator.

The auxiliary omnidirectional antenna is applied in this circuit for the frequency conversion of signal into the stable frequency, determined by the heterodyne, stabilized by quartz. Voltages from commutator and from auxiliary antenna enter the independent high-frequency amplifiers (UVCh) and mixers (Sm. 1), supplied from common/general/total heterodyne.

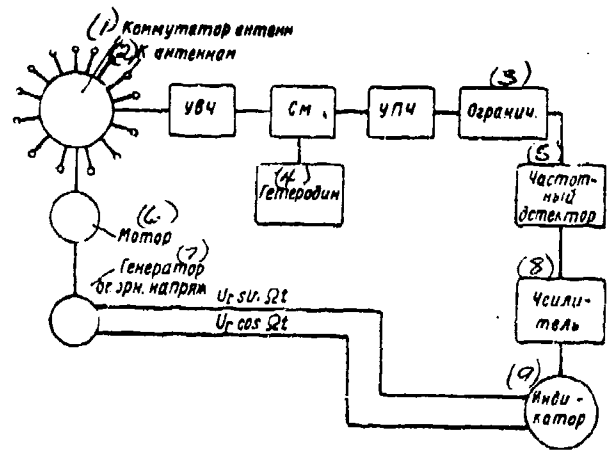

Fig. 8.54. Blo:k liagram of quasi-Doppler radio direction finder.

Key: (1). Commutator of antennas (2). To antennas. (3). Limit. (4). Heterodyne. (5). The FM discriminator. (6). Motor. (7). Generator is supporting/reference. voltage. (8). numerator. (9). Indicator.

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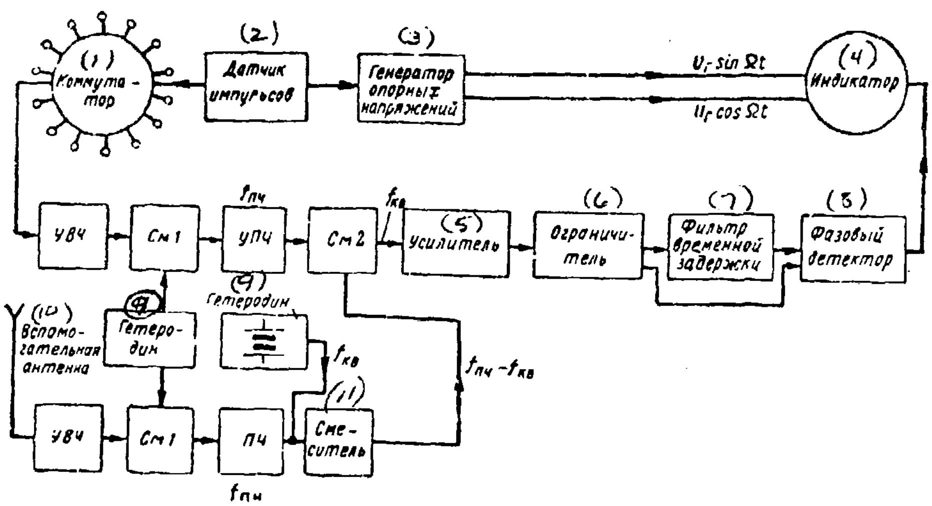

Fig. 8.55. Block diagram of differential-phase radio direction finder.

Key: (1). Commutator. (2). Impulser. (3). Reference generator. (4). Indicator. (5). Amplifier. (6). Limiter. (7). Filter of time delay. (8). The phase discriminator. (9). Heterodyne. (10). Auxiliary

antenna. (11). Mixer.

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The obtained voltages of intermediate frequency separately are amplified in IF amplifiers (UPCh) and (PCh). The voltage of the intermediate frequency of the channel of auxiliary antenna is mixed in the mixer with the voltage of the quartz heterodyne of frequency $f_{кв}$. As a result of transformation, is obtained the frequency $f_{пч}-f_{кв}$. The voltage of this frequency is mixed in second mixer (SM2) with the output voltage UPCh of fundamental channel, forming voltage of frequency $f_{пч}-(f_{пч}-f_{кв})=f_{кв}$. Thus, output potential of second mixer has the stable frequency, equal to the frequency of quartz heterodyne and independent of the frequency of signal and frequencies of the first heterodyne, and also from their possible changes.

After additional voltage amplification of signal undergoes amplitude limitation and is fed to the phase discriminator directly, also, through the filter of time delay. In this filter the signal is delayed to the time $\tau = 2\pi/n\Omega$, equal to time of the switching on of one antenna. The phase discriminator develops a voltage of the frequency of commutation. The latter is compared with respect to the phase with reference voltages in to indicator it gives directly the

value of bearing.

8.9. Automation of removal and averaging of bearing.

Until now, we examined the radio direction finders in which the bearing is counted off by operator. With direction finding the operator takes either one reading, averaged for time of observation or he record/writes several readings and is designed of them average bearing.

Were develop/processed also the methods of automatic, without participation operator, the removal of the readings of bearing and their averaging [8.31, 8.32].

The block diagram of radio direction finder with automatic removal and the averaging of readings is shown in Fig. 8.56. The converter of the output voltage of receiving indicator converts the voltage of receiving indicator so that it becomes convenient for the calculation of the angle of bearing. Under the action of the signals of the circuit of control according to predetermined program, are open/disclosed the circuits of counter-summator of the values of bearings and counter of number of readings, and on termination of

averaging time is switched on resolver, which develops average bearing as quotient of the division of counter readouts.

[Page 511] Instead of the addition directly of readings, it is possible to summarize other parameters from which is determined average bearing. During addition and calculation of average bearing, can be considered the weights of readings, proceeding, for example, from the signal amplitude and character of bearing (ellipticity of image in two-channel radio direction finder, etc.).

Most simply problem is solved in radio compasses with installation on bearing of antenna system or intermediate element between antenna system and receiving indicator (goniometer, phase inverter, the antenna commutator). For the reading of bearing, it is not required to transform the output voltage of receiving indicator, it is necessary to only determine the position of the axis of the rotatable cell/element. It is possible for this to the axis of rotation to mount disk with several concentric path/tracks each of which with the aid of commutator bars or photocells issues the bit of the code of the position of axis. Figures 8.57 shows the location of commutator bars, or the darkened bands on disk, with indication with accuracy 1/64 about (6 bits). On line AA_1, are establish/installed the detachable brushes or the photocells, which record the angle of axis. In Fig. 8.57 angular indication corresponds to 110,101. This

means that the position of axis corresponds to 53/64 circumferences from the beginning of calculation. For the elimination of the errors, connected with the reading of code on the boundaries of sectors, is developed the special code of Gray, and are also proposed other methods [8.30].

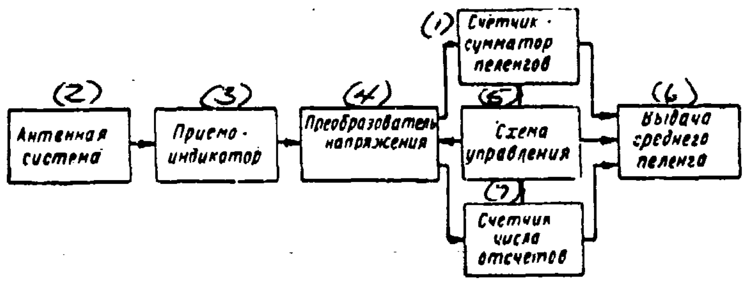

Fig. 8.56. Radio direction finder with automatic removal and averaging of bearings.

Key: (1). Counter is the summator of bearings. (2). Antenna system. (3). Receiver display. (4). Voltage converter. (5). Circuit of control. (6). Delivery of average bearing. (7). Counter of a number of readings.

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Encoders with optical disk are released for a reading to 16 bits (65,536 fixed/recorded points). For obtaining average bearing, must be utilized the latter of four cell/elements of block diagram (Fig. 8.56)

On Fig. 8.58 is given block diagram for taking of bearing in radio direction finder with servo system without the application/use of code disk [8.32].

By the servomotor, controlled by output receiver current, search coil of goniometer is installed on the bearing which usually is counted off for the bearing which usually is counted off on the scale. On the disk, rotated by special motor (25 Hz) and not connected with the axis of goniometer, is establish/installed the photocell which creates current pulse in flip-flop 2, when on it falls light beam from mirror on the axis of goniometer. When the photoelement on disk passes the position of reference point, is closed by relay and flip-flop 1 it creates momentum/impulse/pulse for the beginning of calculation. Time interval between the premise/impulses of flip-flops 1 and 2 is record/fixed by multivibrator and is filled with clock pulses.

The clock pulses, which are created by the gear of motor, are form/shaped with flip-flop 3 and through the modulator fall on counter 1. Simultaneously on counter 2 is checked the number of readings, corresponding to the number of revolutions of disk with photocell.

Resolver issues average bearing on counter 3. In device depicted on Fig. 8.58, pulse separation of calculation (flip-flop 3) corresponds to the displacement of disk to 1°, since time of one turn of disk is equal to 0.72 s. Thus, at the output of device is obtained average bearing in degrees for time of the switching on of servomotor.

Since the mechanism of reading is separated from receiving indicator, readings does not affect the character of transmission.

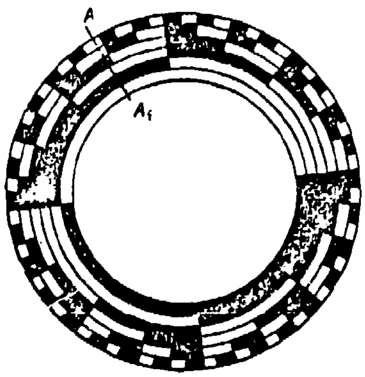

Fig. 8.57. Disk for the reading of bearing.

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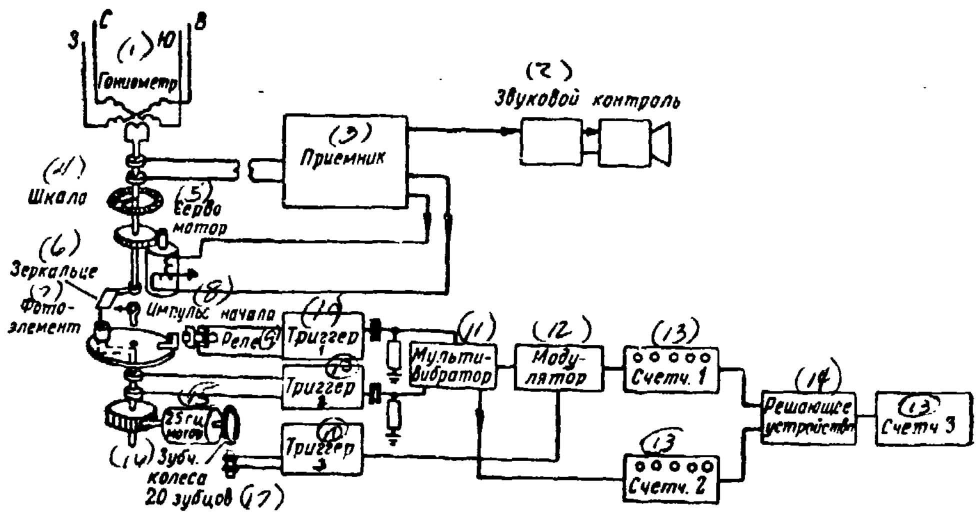

Fig. 8.58. Block diagram of automatic removal of bearing in radio direction finder with servo system.

Key: (1). Goniometer. (2). Acoustic test. (3). Receiver. (4). Scale. (5). Servo-motor. (6). Mirror. (7). Photocell. (8). Momentum/impulse/pulse it began. (9). Relays. (10). Flip-flop. (11). Multivibrator. (12). Counter. (14). Resolver. (15). Hz motor. (16).

Gear wheel (17). teeth.

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In phase-meter radio direction finder the removal of reading is reduced to the measurement of a phase difference between modulating voltage of the frequency of rotation in the output of receiving indicator and the reference voltage. For the measurement of phase difference, it is possible, for example, eacges of the mentioned voltages to create the short-term momentum/impulse/pulses, which correspond to the zero (with increase) instantaneous values of voltages, and to fill time interval between them with clock pulses. For the averaging of bearing, it is possible to use circuit, analogous Fig. 8.56.

In phase-meter radio direction finder with the high-spin motion of acute/sharp radiation pattern, appears the problem of determining the middle of major lobe of the diagram which can be solved by following method. The limiter of receiving indicator isolates two symmetrical points of major lobe. During the passage of the first point, begins the calculation of counter pulses of constant capacitance/capacity (on N of momentum/impulse/pulses). Between the first and second points the period of pulses). Between the first and

second points the period of pulses is equal to T, after the second point they take T/2. Let between points it is containned by n of momentum/impulse/pulses. Counting of n pulses will end after the beginning after the interval of time

$$t=nT+(N-n)\frac{T}{2}=\frac{n}{2}T+N\frac{T}{2}. \qquad (8.76)$$

From formula (8.76) it is evident that the momentum/impulse/pulse, shifted against direction of rotation in the angle, which corresponds count time NT/2, coincides with the middle of the main lobe of radiation. After determining the midlle of major lobe, it is possible to determine single reading and then the averaged bearing.

Figures 8.59 gives the block diagram of the device, which makes it possible to carry out automatic removal and averaging of bearings in two-channel radio direction finder.

The output voltages of two channels of receiving indicator are converted in mixers 1 and 2, moreover the voltage of second channel simultaneously is shift/sheared on phase to 90°, for this into the common/general/total heterodyne before mixer 2, is included the phase inverter, which shifts phase to 90°.

Page 5*5.

Output voltages by mixer $U_1 = E \cos \theta$ and $U_2 = jE \sin \theta$ store/add up and are deducted in sum-and-difference block/module/unit; as a result is obtained the total voltage

$$U_{\Sigma} = E(\cos\theta + j\sin\theta) = Ee^{j\theta}$$

the differential voltage

$$U_{\Delta} = E(\cos\theta - j\sin\theta) = Ee^{-j\theta}.$$

In the measuring unit of a phase difference, is determined the phase difference ψ of stresses U_{Σ} and U_{Δ}, which is equal to the doubled angle of bearing 2θ. Two-place bearing is defined as half of a phase difference ψ.

A phase difference can be measured according to the method, described for phase-meter radio direction finder. By the application/use of supplementary nondirectional antenna it is possible to obtain single-valued bearing. Reading circuit and averaging does not have special feature/peculiarities.

If the output voltages of channels due to the dissimilarity of the amplification of channels will be

$$E_1 = E\cos\theta \quad \text{и} \quad E_2 = aE\sin\theta e^{j\varphi},$$

that

$$U_\Sigma = E(\cos\theta + ja\sin\theta e^{j\varphi}) = E_\Sigma e^{j\psi_1},$$
$$U_\Delta = E(\cos\theta - ja\sin\theta e^{j\varphi}) = E_\Delta e^{j\psi_2},$$

where

$$\mathrm{tg}\,\psi_1 = \frac{a\sin\theta\cos\varphi}{\cos\theta - a\sin\theta\sin\varphi};$$
$$\mathrm{tg}\,\psi_2 = -\frac{a\sin\theta\cos\varphi}{\cos\theta + a\sin\theta\sin\varphi},$$

whence

$$\mathrm{tg}\,\psi = \mathrm{tg}(\psi_1 - \psi_2) = \frac{2a\,\mathrm{tg}\,\theta}{1 - a^2\,\mathrm{tg}^2\theta}\cos\varphi. \qquad (8.77)$$

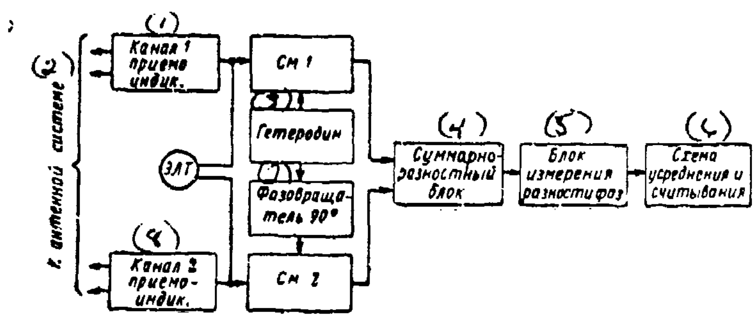

Fig. 8.59. Block diagram of the automatic removal of readings in two-channel radio direction finder.

(1). Channel 1 of receiving indicator. (2). To antenna system. (3). Heterodyne. (4). Sum-difference block/module/unit. (5). Measuring unit of a phase difference. (6). Circuit of averaging and readings. (7). Phase inverter. (8). Channel 2 of receiving indicator.

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From formula (8.77) it is evident that in this case is observed the same error as in usual two-channel radio direction finder [see

(4.9).

Consequently, in a described manner is determined the direction of the transverse of the image of bearing, nonidentity of the amplification of channels giving to by the same errors, as in two-channel radio direction finder.

Phase displacement ε_2 at an angle of 90° - δ (instead of 90°) causes supplementary ellipticity and the same error as nonidentity of the phases in channels. With the image of parallelogram on the cathode-ray tube of two-channel radio direction finder, does not occur the readings of two bearings. Separate readings correspond to the direction of major axis of one of the ellipses, drawn on cathode-ray tube (see 8.3). The averaged reading must correspond to direction of one of the diagonals of parallelogram.

MT-65-58
Principles of Radio Direction Finding,
"Soviet Radio" Publishing House,
Moscow, 1964.
Pages: 517-541 and 636-640.

CHAPTER 9

TESTS OF RADIO DIRECTION FINDERS

List of Designations Appearing in Cyrillic

$k_{\text{гн}}$ = k_{st} = scaling factor

в = fair = fairlead
вых = out = output

и = i = (definition undetermined)

к = comp = compensating

макс = max = maximum

н = load
нс = asym = asymmetric

п = f = field coil

Р, р = loop

Ф = fd = feeder

э = st = standard

В = V = variometer

ГСС = SSG = standard signal generator

З = ground

Preliminary tests of radio direction finders are performed in laboratories, final ones are performed in real operating conditions of the direction finder.

9.1. Laboratory Tests of Direction Finders with a Rotating Loop

Separate parts of the direction finder (the loop, variometers, etc.) require no special tests other than normal ones — measurement of inductance, capacitance, resistance, and coupling coefficient. We shall not dwell here on methods of measurement of these magnitudes.

During laboratory testing of the direction finder as a whole by a generator of standard signals there is required, analogously to normal measurement of receivers, use of an equivalent antenna. A peculiarity of the given case is that receiver-direction finder is fed simultaneously from two antennas: a loop and an open antenna, where the virtual height of the loop changes in a wide range with change of wavelength, and the phase of the emf induced in it differs by 90° from the phase of the emf in the antenna. Furthermore, ordinary generators of standard signals have an asymmetric output (one pole usually is grounded). Connection of output terminals of the generator to the loop creates a symmetry of its circuit, which may not correspond to real operating conditions of the loop.

In Fig. 9.1 there is presented the circuit of the equivalent of the antenna and the loop, considering these peculiarities. Parameters of the circuits are selected in such a manner that $L'_2 + L''_2 \ll L_0$, where L_0 — inductance of the loop; L_a, C_a, R_a, C_{fair} — inductance, capacitance, and resistance of the antenna and capacitance of its fairlead. Under these conditions the receiver has normal load both from the loop and from the antenna.

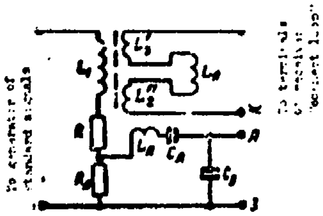

Fig. 9.1. Diagram of equivalent of antenna and loop.

Then we select $R \gg \omega L_1$; then the current through winding L_1 with sufficient accuracy (with error of 1%, if $R > 7\omega L_1$) can be expressed

$$\dot{I}_1 = \frac{E}{R},$$

where E — output voltage of generator.

The emf induced in coils L'_2 and L''_2, will be

$$E'_p = j\omega M I_1 = j\frac{\omega M}{R}E,$$

and voltage on resistance R_a, corresponding to the emf in the antenna, is equal to

$$E'_a = \frac{R_a}{R}E.$$

Coupling between coils L_1 and L'_2-L''_2 is made variable by sine law

$$M = M_{max}\sin\theta.$$

During real work the emf in the frame is

$$E_p = jEh_p\sin\theta,$$

the emf in the antenna is

$$E_a = Eh_a.$$

We equate $E'_a = E_a$, $E'_{loop} = E_{loop}$ and $\alpha E = E$, where α — factor, which is conveniently selected equal to any round number (1, 2, 5,...1/2, 1/5, 1/10, etc.). From this we find

$$\alpha = \frac{\omega M_{\text{макс}}}{h_p R} = \frac{R_a}{h_a R} \quad \text{and} \quad \frac{\omega M_{\text{макс}}}{R_a} = \frac{h_p}{h_a}.$$

Since h_{loop} is proportional to frequency $h_{loop} = \frac{2\pi SN}{\lambda} = \frac{\omega SN}{3\cdot 10^{8}}$, the lase equality is realizable in the whole range of frequencies. From it we find M_{max}, after which, given α, we find R. The reading on the divider dial of the generator of standard signals, multiplied by α, gives field strength in microvolts/meter.

By this circuit we can perform the following tests:

1. Determining sensitivity of the radio direction finder, i.e., the field strength which is required to ensure possibility of direction finding with error not exceeding a given value. For this, there is determined that voltage from the generator of standard signals at which bearing is read with the given accuracy. From the voltage field strength is calculated.

2. Check of exactness of determination of direction. Switching on the direction finder, we find field strength E_1 and E_2 in two positions, corresponding to determination of the direction, with constant output voltage. Depending on the scheme for determining direction these positions can be established either in the receiver itself by turning the variometer, switch, and so forth, or by turning the loop. In the last case in the test circuit turn of the loop is replaced by turn of variometer L'_2-L''_2 from the position corresponding to $+M_{max}$ to position $-M_{max}$. Relation $\frac{E_1}{E_2}$ characterizes exactness of determination of direction.

3. Check of compensation for antenna effects. The problem is to determine the relative emf of the antenna effect which can be compensated. It, obviously, is equal to the maximum emf created by the compensator. To determine this value we determine field strength E_0, creating normal output voltage with the position of the compensator, corresponding to zero emf of compensation. Then we turn variometer L'_2-L''_2 until we obtain zero emf in the loop circuit and the place compensator in the position, giving maximum compensation emf. In this position we again determine field strength E_{comp}, giving the same output voltage. Ratio $\frac{E_0}{E_{comp}}$ gives the value we sought.

4. It is possible to check remaining characteristics of the receiver (selectivity, fidelity, and so forth).

Testing by the two-signal method is accomplished with two equivalents, whose inputs are connected to two generators, and outputs are parallel-connected. Resistances and reactances of the equivalent should be doubled.

Another method of laboratory testing consists of placing the loop in a magnetic field, which is created by current in a horizontal rectilinear wire (line) (Fig. 9.2).

This test should be conducted in a shielded chamber, since during tests with the loop connected (and not with its equivalent, as in the preceding method) external interferences hamper tests a great deal. At a certain distance d from the chamber ceiling we stretch a rectilinear wire, which at one end is joined by a shielded cable to the generator of standard signals, and on the other through resistance R to the metal wall of the chamber.

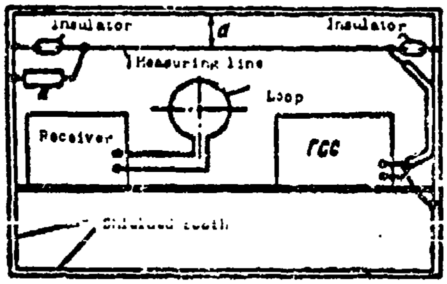

Fig. 9.2. Measuring line for testing a direction finder.

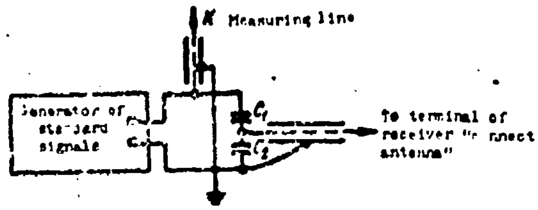

Fig. 9.3. Voltage divider.

The purpose of resistance R is to provide in the wire a traveling wave of current. In traveling wave conditions current in wire, and, consequently, magnetic field strength around it depends little on frequency. Magnetic field strength in these conditions also does not depend strongly on shift of the observation point along the wire.

Under wire there is placed the loop direction finder being tested. With rotation of the loop the minimum emf is induced in it at the time when its plane is perpendicular to the wire.

Magnetic and electrostatic fields of a rectilinear wire at a small distance from this wire do not have as simple a relationship to one another as in the zone of radiation. Therefore, use of the open antenna of a direction finder in its normal position can lead to a relationship of emf's induced in the antenna and loop, absolutely different from the relationship in real conditions. For testing it is necessary to use as the antenna a special section of rectilinear conductor, located in parallels to the test line. The length and distance of this conductor from the

line will be selected in such a way as to ensure a normal relationship of emf's in the antenna and loop. For feed of the antenna circuit it is also possible to use a voltage divider (Fig. 9.3).

First of all it is necessary to select such a resistance R that in the line there is established a traveling wave. Wave impedance of a single-wire line with diameter 2r at distance d from the conducting plane is equal to

$$\rho = 138 \lg \frac{2d}{r}. \qquad (9.1)$$

By this formula there can be found the approximate value of resistance $R = \rho$. Traveling wave conditions in the line are verified by one of the known methods. In this case it is convenient to use the fact that impedance of a line, loaded on wave impedance, is equal to the wave impedance. Due to this, connection to the generator of standard signals of a line, loaded on resistance R, if $R = \rho$, will influence the generator the same as connection of the actual resistance R (will cause the same decrease of its output current). By several tests it is possible to definitize magnitude R, initially found by the formula (9.1). Traveling wave conditions must be verified in the whole range of frequencies of the direction finder.

Line calibration, i.e., determination of the field strength corresponding to the given output voltage of the generator, is produced by a comparator. The antenna of the comparator should be loop-type and of approximately the same dimensions as the loop of the direction finder.

If generator voltage is U, and field strength is E, then $k_{sf} = \frac{E}{U}$ is called the scaling factor, determination of which is the purpose of calibration.

Calibration should be performed at several frequencies within the frequency range of the direction finder. Independence of the scaling factor from frequency is confirmation of the fact that in the line there have been established traveling wave conditions.

It is necessary also to produce calibration for different distances of the center of the loop from the line.

If there is no comparator, calibration can be produced by a loop, whose geometric dimensions are known exactly. The emf on terminals of the loop should be measured by a voltmeter with a very large input impedance. As such voltmeter we use receiver with supply of voltage to the cathode grid of the first tube. The receiver is calibrated from a generator of standard signals.

If maximum emf in the loop is E_{max}, and voltage from the generator is U,

$$k_{rn} = \frac{E_{max}}{h_e U}, \qquad (9.2)$$

where h_e is the calculated effective height of the loop.

For selection of an auxiliary antenna or specifications of the divider feeding the antenna circuit, we should know the effective height of the open antenna of the direction finder h_a.

The emf introduced into the antenna circuit of the direction finder in real conditions is equal to

$$E_a = E h_a.$$

When testing under a line with the help of a divider this emf is equal to

$$E_a = U \frac{C_1}{C_1 + C_2}.$$

From this we find

$$\frac{C_1}{C_1 + C_2} = \frac{E}{U} h_a = k_{rn} h_a. \qquad (9.3)$$

The sum of capacitances $C_1 + C_2$ should be equal to the capacitance of the antenna C_a. Formula (9.3) gives the possibility of determining C_1 and C_2:

$$C_1 = C_a h_a k_{rn}, \qquad (9.4)$$

$$C_2 = C_a (1 - h_a k_{rn}). \qquad (9.5)$$

Testing under a line permits determining the same parameters of a direction finder as testing with the help of an equivalent antenna. Furthermore, testing under a line permits checking the sharpness of minima and the magnitude of errors depending upon frequency, field strength and other factors.

For checking selectivity by the two-signal method there is stretched a second line, perpendicular to the first and fed by a separate generator. Frequency and field strength of the disturbing radio station are established on this second generator.

It is necessary to note that neither the first nor the second method of laboratory testing corresponds fully to real conditions of work and, therefore, they can give results, differing from results of tests in operational conditions. Nonetheless, laboratory tests are very desirable, since thanks to the easy of shifting frequency, change of amplitude of the fed voltage, etc., tests can be conducted more widely and deeply than during tests on real work. Here, there can be revealed defects which would be passed over during performance tests.

Of the two methods described, obviously, the second corresponds more closely to real conditions of work of the direction finder, but carrying it out is somewhat more complicated than for the first.

9.2. Laboratory Tests of Direction Finders of a Goniometric System

Tests of Loops

Besides normal checking (determination of inductance, self-capacitance, damping, and so forth) for loops of goniometric systems it is very important to check the magnitude of mutual inductance between them. Absence of mutual inductance simultaneously confirms their mutual perpendicularity. From smallness of permissible magnitude of mutual inductance (permissible coupling coefficient is of the order of 0.3-0.4%) normal bridge and resonance methods are insufficiently exact.

A measuring circuit, permitting a reading, with the required degree of accuracy, is presented in Fig. 9.4. B is a variometer with very small inductances of windings (considerably smaller than inductance of loops), but with a fairly strong maximum coupling between them (K = 0.4 to 0.6). The high coupling coefficient permits sufficiently accurate calibration of the variometer. One of the windings of the variometer, series-connected with one of the loops, is fed from the generator; the other winding of the variometer and the second loop are also coupled in series and are joined to the cathode grid of the first tube of the receiver. Audibility on the receiver output turns into zero when the coefficient of mutual inductance of the variometer is selected equal to the coefficient of mutual inductance of the loops.

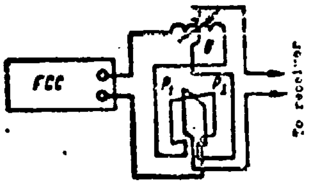

Fig. 9.4. Measuring circuit of small mutual inductance.

The generator and receiver, and also the variometer must be shielded, and all wiring is carried out in such a way as to exclude spurious couplings between circuits of the two loops.

Testing of the Goniometer

In the goniometer all its electrical parameters — inductances and distributed capacitances of all coils and maximum coupling coefficient between each of the field and the searcher coils — are to be checked. It is necessary also to check the coefficient of mutual inductance between the two field coils. This measurement can be made by the same scheme as analogous measurement for loops.

The most important test of a goniometer is determination of the error curve. Measurement of erros can be taken at high and low frequencies.

For checking at high frequency we compare the tested goniometer with a standard one, connecting them as shown in Fig. 9.5. Let the rotor of the standard goniometer turn about the first coil of the stator at angle θ_{st}. Assuming that the standard goniometer is absolutely exact we can present the emf's induced in stator coils in the form

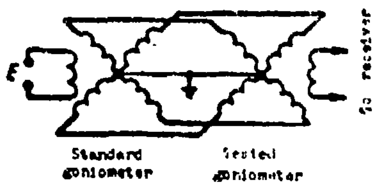

Fig. 9.5. Comparison of goniometer with a standard one.

$$E_1 = j\frac{E\omega M_1}{Z_1}\cos\theta_s,$$
$$E_2 = j\frac{E\omega M_1}{Z_1}\sin\theta_s,$$

where M_1 — maximum mutual inductance;

Z_1 — impedance of rotor of standard goniometer;

E — feed voltage.

Currents in stator coils will be

$$I_1 = jE\frac{\omega M_1}{Z_1(Z_s+Z_{x1})}\cos\theta_s,$$
$$I_2 = jE\frac{\omega M_1}{Z_1(Z_s+Z_{x2})}\sin\theta_s,$$

where Z_{st}, Z_{x1} and Z_{x2} — impedances of stator coils of the standard and investigated goniometers.

Normally impedances of two stator coils are equal to one another, i.e.,

$$Z_{x1} = Z_{x2} = Z_x.$$

If the searcher of the tested goniometer is turned an angle θ_x, then the emf induced in it will be

$$E_4 = -E\frac{\omega^2 M_1 M_2}{(Z_s+Z_x)Z_1}\cos(\theta_s-\theta_x),$$

where M_2 — maximum mutual inductance between field and searcher coils of the tested goniometer.

This emf turns into zero when $\theta_x = \theta_{st} + 90^\circ$. Thus, setting the rotor of the standard goniometer at some angle θ_{st}, we should obtain disappearance of audibility upon setting the rotor of the tested goniometer at an angle $\theta_{st} + 90^\circ$. The difference between this angle and the angle of setting, at which we obtain real disappearance of audibility, directly gives error of the goniometer. In an analogous way we can test a goniometer with three or four field coils.

The circuit of other method of testing at high frequency is presented in Fig. 9.6. If we select resistance so that $R_4 \ll \omega L_f$, where L_f — inductance of field coil,

R_4 — impedance of the divider, voltage division depends exclusively on the magnitude of resistances. Thus, voltage on one of the field coils will be

$$E_1 = E\frac{R_1}{R_4},$$

and on the other field coil,

$$E_2 = E\frac{R_3}{R_4}$$

etc., where R_1, R_2, ..., R_4 — resistance from beginning of divider to the corresponding tap.

Analogously to the preceding we find the emf in the searcher coil, assuming the goniometer is free from errors. Thus, with coupling of field coils into taps R_1 and R_3 we obtain

$$E_s = E_1\frac{M}{L_0}\cos\theta + E_2\frac{M}{L_0}\sin\theta = \\ = E\frac{M}{L_0}\left(\frac{R_1}{R_4}\cos\theta + \frac{R_3}{R_4}\sin\theta\right),$$

where M — maximum mutual inductance of the field and searcher coils of the goniometer.

The emf turns into zero when $\tan\theta = -\frac{R_1}{R_3}$.

If the goniometer gives error, then disappearance of audibility will occur at another angle Φ. Error of the goniometer will be equal to

$$\Delta = \Phi - \theta.$$

Thus, attaching the ends of field coils to various terminals of the divider and determining the position of the searcher corresponding to vanishing of audibility in the telephone, we can determine the error of the goniometer at different angles. It is possible to have a comparatively small number of taps in the divider (3-4), in order to obtain sufficiently closely located points for construction of the error curve.

Shielding of the generator, receiver, and divider, thoroughness of location of wiring in this method are as necessary as in the method of a standard goniometer.

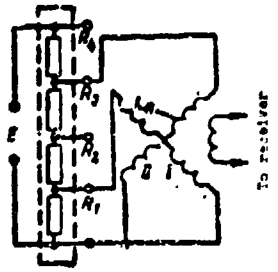

Fig. 9.6. Testing a goniometer by a divider.

The divider itself should be made inductionless and non-capacitive, possess a small skin effect, which is necessary for preservation of constancy of the ratio of resistances during change of frequency. One should make it with the same care as, e.g., attenuators of generators of standard signals.

The circuit for checking a goniometer at low frequency is shown in Fig. 9.7. In it R_1 and R_2 —

precision resistance boxes; L_1 and L_2 — field coils of goniometer; L_3 — its searcher coil; T — telephone (low-resistance).

The circuit is fed from an af generator G. Resistances R_1 and R_2 should be taken considerably larger than induced resistance of field coils with frequency of measurement ω, i.e., $R_1 >> \omega L_1$ and $R_2 >> \omega L_2$. In this case currents I_1 and I_2 are determined by equalities

$$I_1 = \frac{E}{R_1} \text{ and } I_2 = \frac{E}{R_2}.$$

If the goniometer was made absolutely exactly, the emf induced in the searcher coil would be

$$E_3 = I_1 \omega M \sin \theta \pm I_2 \omega M \cos \theta.$$

Rotating the searcher coil until audibility disappears in the telephone, we obtain angle θ from equation

$$I_1 \omega M \sin \theta \pm I_2 \omega M \cos \theta = 0$$

or

$$\operatorname{tg} \theta = \pm \frac{R_1}{R_2}. \qquad (9.6)$$

If the goniometer has error, audibility will disappear at another angle $\Phi = \theta + \Delta$, where Δ — degree of error. The method of checking consists in establishing ratio $\frac{R_1}{R_2}$ conforming to angles $\theta = 0^\circ$, 10°, 20°, etc., and determining angle Φ at which sound disappears in the telephone. Difference

$$\Phi - \operatorname{arc\,tg} \frac{R_1}{R_2} = \Delta$$

directly gives error of the goniometer. To each ratio $\frac{R_1}{R_2}$ there correspond two angles differing approximately by 180°, at which audibility vanishes.

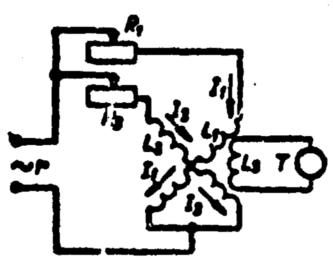

Fig. 9.7. Circuit for checking of goniometer at low frequency.

Thus, the goniometer is checked from 0° to 90° and from 180° to 270°. To check the second half of the dial the ends of one of the field coils are connected, which corresponds to a change of sign in formula (9.6). During work it is necessary to watch to see that the generator does not directly influence the searcher coil, and that current in the telephone does not influence the field coils.

Analogous circuits can be easily composed for testing multiwinding goniometers.

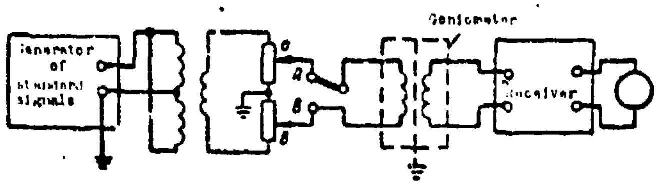

Fig. 9.8. Diagram for checking symmetry of goniometer.

So that in the goniometric system there is no antenna effect, it is necessary to ensure complete symmetry of field coils of the direction finder. Check of symmetry of the goniometer can be performed by the circuit in Fig. 9.8. Voltage from the generator of standard signals is brought to the field coil through a symmetric transformer (see § 4.3) and a potentiometric circuit of resistances.

The searcher coil is connected to the receiver. In switch position A voltage acts between ends of the field coil, which corresponds to reception of a two-phase wave. In switch position B the emf acts between both ends of the field coil and the "ground" (i.e., the frame of the goniometer, cathode of the first tube of the receiver and its frame), which corresponds to reception of a single-phase wave. A completely symmetric goniometer in the second switch position will not transmit voltage to the searcher coil.

In practice measurement is performed in the following way. Setting the switch in position A, tuning the receiver and turning the searcher coil to the position of maximum coupling with the tested field coil, we regulate the voltage of the generator of standard signals to obtain a conveniently read receiver output voltage U. Let us assume that here the voltage of the generator of standard signals is equal to E_1.

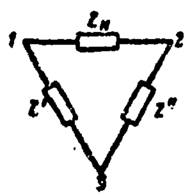

Fig. 9.9. Circuit of asymmetric loading.

Then we shift the switch to position B, increase output voltage of the generator of standard signals and turn the searcher coil to obtain maximum receiver output voltage. Let us assume that voltage of the generator of standard signals, necessary for production of the same receiver output voltage U, in this case is equal to E_2. Then the relative degree of asymmetry of the goniometer is characterized by ratio $\frac{E_1}{E_2}$.

In carrying out tests it is necessary to ensure symmetry of the transformer, equality of potentials at points a and b, and also to avoid any asymmetry of the

goniometer (for instance, because of asymmetric position of wires to the switch).

Measurement of asymmetry by another method is carried out with the help of an hf resistance bridge. Asymmetry is caused by unequalness of capacitive or in general, any impedances between terminals 1 and 2 of load Z_{load} (in this case the goniometer) and the ground. In Fig. 9.9 these impedances are designated Z' and Z''. The asymmetry parameter is equal to

$$k_{as} = \mathrm{mod}\,\frac{Z' - Z''}{Z' + Z''}.$$

We take three measurements of admittances:

1) between point 1 and grounded point 2 (Y_1);
2) between grounded point 1 and point 2 (Y_2);
3) between short-circuited terminals 1 and 2 and ground (Y_3);

$$Y_1 = \frac{Z' + Z_a}{Z' Z_a};\quad Y_2 = \frac{Z'' + Z_a}{Z'' Z_a};\quad Y_3 = \frac{Z' + Z''}{Z' Z''}.$$

It is easy to see that

$$k_{as} = \mathrm{mod}\,\frac{Y_2 - Y_1}{Y_3}.$$

Both methods of measurement of asymmetry are applicable also to measurement of asymmetry of the input of the receiver and of other elements.

Test of a Radio Direction Finder as a Whole

To test a loop radio direction finder of a goniometric system in laboratory conditions there can be employed the same two methods as for testing a direction finders with a rotating loop, i.e., testing with an equivalent and testing with the help of a line. The equivalent presented in Fig. 9.1 gives the possibility of feeding emf only to one of the field coils. The remaining field coils of the goniometer should be closed to the same equivalents with closed input terminals.

When testing by the two-signal method there can be used a second field coil, to which there is fed an emf from a second generator through an antenna equivalent.

In the case of an external system of two spaced antennas it is also possible to compose an equivalent. Its circuit is presented in Fig. 9.10. Here C_a, C_{fd}, C, C_{fair} — capacitances of spaced antennas, feeder, auxiliary antenna and its fair-lead; L and L_a — inductances of the auxiliary and the spaced antennas. Selection of magnitudes R, R_a and M is analogous to the preceding case. It should be stressed that testing by an equivalent has meaning only for those systems, for which natural waves of antennas considerably differ from working waves.

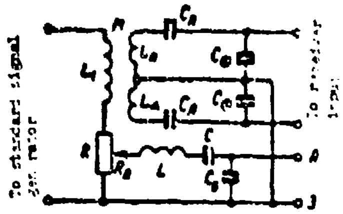

Fig. 9.10. Circuit of the equivalent for a system of speed antennas.

9.3. Laboratory Tests of Radio Direction Finders with Wide Antenna Spacing

Separate component parts of the radio direction finder (hf transformers, time delay line, switching circuits, indicators, etc.) are checked by usual methods.

The antenna system of a radio direction finder with wide spacing antennas consists of a large number of antennas, in which there are induced emf's of identical amplitude, but with different phases in accordance with geometric location of the antennas (§ 3.11). Antennas are connected to an antenna switch or a reception-indicator.

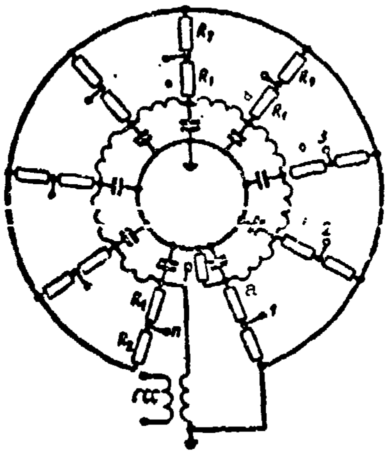

Fig. 9.11. Circuit of the equivalent of a circular antenna system with wide spacing.

During laboratory tests of a radio direction finder it is necessary to be able to introduce to inputs of the antenna switch (or reception-indicator voltage of identical amplitude, the phase of which varies by a given law. For this there is used a special antenna equivalent. In Fig. 9.11 there is presented the circuit of the equivalent for laboratory testing of a radio direction finder with a circular antenna system. It consists of a natural or artificial long line, fed by a generator of standard signals and loaded on an impedance, equal to wave impedance. The section of long line is designed in such a manner that on terminals of the long line of the equivalent a, b, c, etc., voltages have identical amplitudes and phases, equal to phases of the emf of corresponding antennas.

Phases of voltages are calculated for the case when there is produced reception of a radio station from a definite direction.

Between terminals of the long line and the ground there are coupled resistance R_1-R_2 such magnitude that $R_1 \gg R_2$ and $R_2 = \rho_{fd}$, - where ρ_{fd} wave impedance of feeders leading into the antenna switch (matched loading of feeders from the antennas is assumed).

Thus, on the output terminals of the equivalent 1, 2, 3, ..., n there are voltages of identical amplitude with phases, corresponding to phases of emf antennas. Output resistances on these terminals are equal to ρ_{fd}. Application of decoupling resistances R_1 removes influence of loads of the antenna switch on the amplitude and phase of voltages at points 1, 2, 3, ..., n.

The antenna equivalent permits checking the overall efficiency of equipment, determining instrument accuracy for fixed directions and sensitivity. Instrument accuracy is determined connecting the output of the equivalent 1, 2, 3, ..., n first to terminals 1', 2', 3', ..., n' of the antenna switch. Here, on the antenna switch during direction finding there should be read an angle, corresponding to that bearing, for which the long line is calcule '. Then the output of equivalent 1, 2, 3, ..., n are switched to terminals 2', 3', ..., n', 1' of the antenna switch, 3', 4', ..., n', 1', 2', etc. Each switching corresponds to displacement of the direction of bearing an angle, equal to the angle between the antennas. The difference between readings on the bearing indicator of the radio direction finder and calculated bearings corresponds to instrument errors. To determine sensitivity it is necessary to preliminarily find coefficient κ of transmission of voltage from input terminals of the equivalent to its output terminals 1, 2, 3, ..., n with connected loads.

If to the equivalent's input there is fed voltage U, then $E = \frac{U}{kh_{eff}}$, where h_{eff} is the effective height of the antenna.

With an unmatched loading of feeders instead of ρ_{fd} it is necessary to couple in at each frequency its own Z_a, corresponding to input impedance of the antenna and feeder together.

By this method we also determine the directivity pattern of the antenna system.

9.4. Tests of Direction Finders in Real Conditions of Work

When testing a direction finder in the place of installation it is necessary to check separate parts of the antenna-feeder device (single angennas, feeders, etc.) and correctness of their geometric location. Tests are performed by methods, described in [9.3].

Tests of a radio direction finder have the goal of determining: instrument error of the direction finder, magnitude and nature of local errors, general accuracy of the direction finder, general sensitivity of the direction finder, the characteristic and coefficient of directivity of its antenna system.

Determining Instrument Error of a Radio Direction Finder

It is not possible to determine instrument error for all systems of direction finders. Thus, direct determination of instrument errors for direction finders with a fixed outdoor system is impossible, if the latter is too bulky. In these cases it is necessary to be limited to analysis of separate sources of instrument error on the basis of laboratory tests and tests, which are described in the following point.

Instrument error is most exactly and simply determined for goniometric direction finders, for which structure and dimensions of the outdoor equipment are such as permit rotation of it in the process of testing. For this purpose the external device of the direction finder is set on a special machine, permitting us to turn it at known angles. Tuning to some station, by rotation of the goniometer we find its bearing. We then turn the outdoor system a certain angle (for instance, 10-15°) and repeat fixing. The new reading on the goniometer should differ from the first by the angle of rotation of the outdoor system. Performing such tests for several angles from 0° to 360° and at various frequencies, we can obtain a sufficiently full judgement of instrument accuracy of the direction finder.

Special difficulties are presented by tests of direction finders with calculation of polarization errors. Thus, to determine standard polarization error one should place the direction finder in an electromagnetic field with a known slope of the wave front and angle of polarization. For creation of such a field a local generator is placed at a considerable height (on a mast, ballon, etc.) and is equipped with a radiating dipole, which is set at such an angle as creates a field with the necessary turn of the plane of polarization.

Distance from the direction finder to the generator should be sufficiently great. For direction finders of short waves this distance is practically of the order of 100 m or more. To create an angle of incidence of 45°, corresponding to conditions of test of the error of a standard wave, height of lift of the generator should also be near 100 m. This causes evident practical difficulties, because of which in most cases we are limited to smaller height of rise of the emitter. So that polarization error is not small, the angle of rotation of the plane of polarization γ is made greater than 45°.

If the angle of inclination of the wave front is very small, to satisfy the shown condition angle γ should be close to 90°. Inconvenience of such a condition of tests is the small magnitude of the vertical component of field strength and, consequently, the weak reception power.

To produce reliable results of measurement it is necessary to ensure strict symmetry of the generator, since the presence of a single-cycle wave in a radiating dipole does not permit establishing the exact relationship between magnitudes of vertical and horizontal components of the field. The direction finder should be located in a plane, perpendicular to the plane containing the radiating dipole. Otherwise the ratio of horizontal and vertical components of field strength is not equal to the tangent of the angle of rotation of the dipole. Such a phenomenon does not occur with a radiating loop; therefore in installations for checking polarization errors they chiefly apply a loop as the emitter.

Determining the Magnitude and Nature of Local Errors

To determine errors of the direction finder the generator shifts around it. Direction finding is produced at different positions of the generator, and results are compared with true angles. These angles, depending upon the circumstances, are determined either by visual direction finding, or on the map by known positions of the mobile generator.

The mobile generator is carried, trucked or is placed on a ship, aircraft, etc. If the direction finder is placed on a mobile object (ship, aircraft), one can determine error by a fixed source of radiation by means of shifting the direction finder itself.

The distance from the direction finder to the source of radiation should be sufficiently large, in order to consider the field near the direction finder as the field of radiation, i.e., the distance should exceed the wavelength. Furthermore, the distance should be sufficient, that there is formed an approximately flat wave front:

$$r > \frac{2(2b)^2}{\lambda},$$

where 2b — antenna aperture (spacing of vertical antennas).

The generator should be located in an open locality far from objects which could create a field of secondary radiation.

The error curve, found by the shown method, contains both instrument errors of the radio direction finder, and also locality errors. In order to separate errors, it is necessary to determine instrument errors by other methods (see the first point of this paragraph) or conduct inspection of local errors by a radio direction finder, whose instrument errors are small and are known. In separate cases judgement about whether error is local or instrument is facilitated by consideration of the dependence

of error on the distance to the radiator, i.e., from shift of the heterodyne with constant azimuth. Instrument error does not depend on range, and local error changes with change of range.

With a separate re-emitter the dependence of error, caused by it, on the distance between the heterodyne and direction finder has a regular sinusoidal nature, which facilitates detection of such error.

The general character of the error curve also depends on the distance between the direction finder and generator (§ 10.5). When this distance is small (for instance, 100-300 m), influence of locale and surroundings is not transmitted completely in the obtained error curve, since relative magnitudes and phases of field strength of reverse emitters differ from corresponding magnitudes obtained under the influence of a wave from a very distant source.

It is possible to recommend such tests mainly for checking instrument errors under the condition that local errors are minute.

With large distances (for instance, 3-5 km) conditions of testing are closer to real conditions of work of the radio direction finder and more fully reflect local errors, including and the influence of distant surroundings. These tests should be conducted during the introduction of the radio direction finder into service to judge its general accuracy.

It is necessary to note that the curve of local errors, taken with the help of a generator located on the surface of the earth, can not preserve its form during reception of reflected rays, since reverse radiations, provoking local errors, change their character under the influence of an abnormally-polarized field.

Determining General Accuracy of a Radio Direction Finder

Because of the above-indicated difficulties separate determination of instrument errors, although of considerable interest, cannot completely characterize general accuracy of a radio direction finder. On the other hand, determination only of the general accuracy of a direction finder, the most important of its operational characteristics, says little to the designer, since he cannot find which part of errors can be eliminated by improving the design of the instrument and which part can be eliminated by improving its location. Therefore, whenever possible, one should make a full investigation of the direction finder, determining both instrument and local errors, and also the general accuracy of the direction finder.

To determine general accuracy of a direction finder we perform direction finding

of different known objects, located as far as possible in different directions, at different distances and working on different waves.

Comparing the found radio bearing with the true one, determined by calculation or plotting on a map, we find error. For the characteristic of general accuracy of a direction finder we find the mean error. According to the theory of errors it is most correct to take for mean error the mean quadratic error, but for simplification of calculations we frequently limit ourselves to finding the arithmetic mean error (without taking into account sign).

For radio direction finders, working on short and medium waves, during direction finding at night scattering of errors is so great that their average magnitude insufficiently characterizes work of the direction finder. Methods of analysis of observations are presented in Chapter 11.

Determining General Sensitivity of a Radio Direction Finder

To determine general sensitivity of a direction finder we simultaneously perform direction finding and determination of field strength by a comparator. During direction finding by hearing we note the angle of silence. During visual direction finding we note the line width of the image of bearing or the magnitude of oscillations of the pointer of the indicator from the influence of noises.

On ultrashort, medium and long waves sensitivity can be determined by means of observation of different remote radio stations. On short waves observation of remote stations because of the influence of fade-outs leads to very inaccurate determinations of sensitivity. Therefore, on short waves sensitivity is better determined by using a local generator, removed a certain distance from the direction finder.

Field strength in all cases is measured by a comparator. If there is no comparator, sensitivity can also be determined by the local generator, equipped with an ammeter in the antenna, whose virtual height is known. In this case field strength near the generator is calculated by the corresponding formulas. This method is less exact than direct comparison.

Sensitivity of a radio direction finder may also be found by means of calculation, if there are determined the sensitivity of the receiving equipment and the virtual height of the antenna system (see §§ 2.7-2.11).

One can determine virtual height in the following manner (Fig. 9.12): at a certain distance d from direction finder DF there is located emitter G. Exactly the same distance d from the latter there is placed comparator C, so that field strength

for the direction finder and for the comparator is identical. The distance between the direction finder and the comparator should be such as to exclude their interaction. Field strength created by the emitter is measured by the comparator. Its value E_0 should substantially exceed the level of field strength of the external interferences. The direction finder is tuned to the wave of the emitter, and its antenna system or goniometer is set in a position, corresponding to maximum reception power. Voltage on its output is measured by a voltmeter. By gain control instruments output voltage is set such that work of the receiver in the linear region is ensured. To ensure linearity there will also be turned off the automatic gain control.

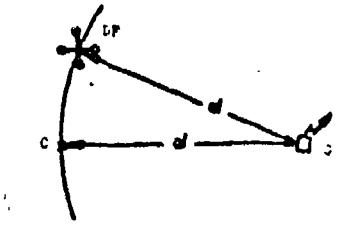

Fig. 9.12. Diagram of location of instruments during measurement of virtual height.

Then we disconnect the antenna system from the receiver and connect to it a generator of standard signals through antenna equivalent. In the goniometer system all non-operating windings of the goniometer should here be loaded on antenna equivalents Eq (Fig. 9.13). Without touching receiver controls, we adjust the voltage of the generator of standard signals to obtain on the receiver output the same voltage as was obtained earlier from the external emitter. Obviously, in this case voltage in the antenna circuit E_g from the generator of standard signals is equal to the voltage in the antenna circuit E_0H_a, which the external emitter created.

On the basis of this we can find virtual height as $H_a = \frac{E_g}{E_0}$*.

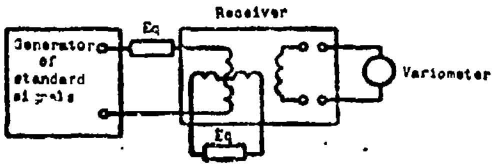

Fig. 9.13. Circuit diagram of receiver.

Determining the Directivity Pattern and the Directivity Factor

To determine the directivity pattern of the antenna system of the direction finder in the horizontal plane it is necessary to determine the dependence of output voltage of receiver on angle, which the direction of propagation of the wave forms

*This formula is valid when resistance of the equivalent is equal to resistance of the antenna.

with the plane of the antenna system. In a system with a turning antenna this angle can be changed by rotation of the antenna system itself. The radiation pattern of a radio direction finder with fixed antennas can be found by rotating the device intended for rotating the radiation pattern (goniometer, commutator) and noting the dependence of output voltage on the angle of rotation of the goniometer or commutator. Another method of finding the radiation pattern consists moving the sender around a fixed antenna system on a circumference, whose center coincides with the center of the antenna system.

One should equip the sender with an ammeter for measuring in it the current, which should be maintained strictly constant. To ensure great accuracy it is possible to recommend monitoring constancy of field strength in the direction finder by a comparator.

Angular shift of the sender is measured by a compass, theodolite, or another such instrument.

Output voltage of the receiver is measured by a voltmeter. It is necessary beforehand to check the range of voltages, in which output voltage linearly depends on input, and during the whole test voltage should not leave the linear region.

Dependency $\frac{U_{out}}{U_{max.out}} = F(\theta)$, presented graphically in polar or cartesian coordinates, is the sought directivity pattern. The directivity factor in the horizontal plane is determined graphically from the directivity pattern:

$$D_г = \frac{2\pi}{\int_0^{2\pi} \rho^2 d\theta},$$

where

$$\rho = \frac{U_{вых}}{U_{вых\ макс}} = F(\theta).$$

Thus, to determine the directivity factor it is necessary to construct the curve of the dependence of ρ^2 on angle θ and by planimetry find its area $\int_0^{2\pi} \rho^2 d\theta$. Number 2π, divided by the found area, gives the directivity factor.

If field strength is measured by a comparator, then before completing the shown calculations it is necessary to introduce a correction in values of measured output voltages, determining magnitude

$$U'_{вых} = U_{вых} \frac{E_{макс}}{E},$$

where E_{max} — maximum field strength;

E — field strength, measured at the same position of generator, at which U_{out} is measured.

During construction of the radiation pattern we use values of U'_{out} and $U'_{max.out}$.

Measurement of directivity patterns in the vertical plane causes considerable technical difficulties; with this purpose it is possible to use aircraft and helicopters [9.3].

NT-65-58
Principles of Radio Direction Finding,
"Soviet Radio" Publishing House,
Moscow, 1964.
Pages: 542-560.

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CHAPTER 10

DIFFERENT APPLICATIONS OF RADIO DIRECTION FINDERS

List of Designations Appearing in Cyrillic

и = i = [definition undetermined]

макс = max = maximum

пр = long = longitudinal

р = loop

П = RDF = Radio Directional Finder

п = trans = transverse

Below are certain practical instructions on selecting the site, installation, and adjustment of ship, aircraft and ground radio direction finders. In this chapter we used materials of published manuals and recommendations [1.10, 10.1, 10.6, 10.7].

10.1. Ship Radio Direction Finder

Selection of Site

For installation of the antenna array of a ship radio direction finder one should select the place, most removed from metallic parts of the ship. Therefore, the antenna array should be assembled as high as possible above the hull of the ship and as far as possible from stacks, masts, antennas and metal superstructures.

To indicate beforehand the best place is difficult. At medium waves the ships hull usually is the main influence. Knowing dimensions of the ship, it is possible by the formulas (5.58), (5.60) to approximately calculate deviation, caused by the hull when placing the antenna array directly on the deck or on a mast.

On short waves of greatest influence are antenna-like objects (masts, stacks, and so forth), tuned in resonance with the frequency of direction finding, when on their length there is laid-off a quarter of the wavelength or three quarters of the wavelength (Fig. 5.14).

Thus, 30 m mast creates the greatest deviation at frequency $f = 2.5$ Mc; the captain's bridge 15 m high, at frequency $f = 5$ Mc; a 11 m stack, at frequency $f = 7$ Mc. Range of frequencies at which such objects act depends on their transverse dimension. An antenna made from a conductor practically has effect in a range of ±5% on both sides of resonance frequency; a stack, in range ±10%; a bridge, in range approximately ±30%, etc.

On ultrashort waves large structures shield reception, and direction finding becomes impossible.

It is necessary before inspection on a ship to consider on the plans of equipment of ship the most convenient places and to calculate for them on the basis of materials of Chapter 5 maximum deviations.

Antenna systems with spaced loops or with spacing of antennas greater than λ, because of their somewhat sharper radiation pattern, are less subject to the influence of fields of reverse radiation. An antenna array with spaced antennas with small spacing reverse emitters is influenced to a still larger degree than a loop array, because of stronger action of the electrostatic fields on vertical antennas.

After places for mounting of the antenna system are noted (usually several), it is advisable to test them with a portable radio direction finder, if possible, and select the best place. To find the fitness of the place we take the curve of deviation along by one of the methods shown below. The best place will be the one, for which maximum deviation is less in the working range of frequencies of the radio direction finder. On medium waves the curve of deviation should be very close to quadratic (quadratic deviation is easy to compensate) and vary little with frequency.

On short and ultrashort waves deviation should vary smoothly with change of relative bearing of the station and with change of frequency. Furthermore, there should be obtained complete angles of silence for bearings of all directions 0-360° in a sound radio direction finder, small ellipses of images (not more than 15%) in an automatic two-channel radio direction finder, and precise readings of bearing in other systems of direction finders.

Finding of the curve of deviation and check of the quality of direction finding should be done at sea. As long as ships are in the dockyard, it is impossible to do these jobs, since around ship there usually are many foreign objects, creating additional errors.

The place for mounting receiver equipment should be convenient for work. The length of the high-frequency cable required for connection of the antenna array (of goniometric or other type) with the receiver should be as small as possible. The receiver equipment of a radio direction finder should be installed in the chart house, since the navigatory uses the radio direction finder.

Mounting of the Antenna Array of a Radio Direction Finder

Let us consider cases, when in the ship radio direction finder as the antenna array there are applied a rotatable loop, a goniometric system of two

mutually-perpendicular loops, a rotatable system of spaced loops, and a system with spaced antennas.

The antenna array is installed in such a manner that its axis of symmetry lies in the diametrical plane of the ship; otherwise there may appear a coefficient of constant deviation A, varying with frequency. Such deviation is difficult to compensate.

An auxiliary antenna, if it is not provided in the construction of the antenna array, is taken as far as possible vertical and located in direct proximity to the directional antenna.

In a rotatable system with reading by the minimum the dial is oriented in such a manner that it reads angle 90-270° when the plane of the system coincides with the diametrical plane of ship, or 0-180° when the plane of the system is perpendicular to the diametrical plane. In a goniometric system one loop or pair of antennas are usually mounted in the diametrical plane; the second is mounted across the ship. With unequal dimensions of loops (for a medium-wave radio direction finder) the smaller loop is installed along the longitudinal axis, thanks to which quarter deviation partially is compensated.

For a goniometric radio direction finder it is necessary to check correctness of connection of the ends of field coils of the goniometer to the loop device and of ends of the searcher coil to the receiver.

For this we listen to and fix any radio station, when to goniometer there is joined only one longitudinal loop (transverse is disconnected); the bearing should be 0° or 180°. If instead of 0° or 180° the bearing is equal to 90° or 270°, then to the longitudinal loop we join the other field coil of the goniometer.

Further we fix a radio station, when one transverse loop is connected; bearing should be 90° or 270°.

Finally, we join both field coils and check matching of fields. With rotation of the ship counterclockwise the bearing should increase; if this is not so, one should exchange ends in any of the field coils.

Correctness of determination of direction is attained by true connection of the ends of searching coil. If direction is determined incorrectly, we switch the ends of the searcher coil.

In the same manner connection to the receiver-indicator of a two-channel radio direction finder is checked.

It is very important that near the outdoor equipment of a radio direction finder all touching metallic parts (for instance, bulkhead of the bridge, guard rail, struts etc.) have good contacts with the hull; otherwise with variable contacts, there is obtained variable deviation. Furthermore, during disturbance of contacts there is audible a crackling in the telephone of the receiver.

Cables and other metallic rigging nearest to the antenna array should have a length, smaller than $1/4\ \lambda_{min}$. If the antenna array is mounted on a separate mast, the upper part of the mast should be free of yards standing out to the sides, and so forth. Guys for the mast should be symmetrically located about the antenna system.

Taking the Curve of Deviation of a Ship Radio Direction Finder

The bearing q of a radio station, lying at heading p to the longitudinal axis of the ship, under the influence of metallic objects located around the loop (antennas, guys, metal hull, and so forth), is incorrect.

Deviation f is equal to

$$f = p - q.$$

The formula for f gives the absolute value and sign of deviation. Deviation generally varies with heading and depends on the length of the fixed wave. After installation of a radio direction finder before using it, it is necessary to determine deviation for all directions from 0° to 360° and for needed waves (take the curve of deviation). During subsequent work on the radio direction finder we use curves of deviation to determine corrections to bearings.

The curve of deviation is taken chiefly by a visible radio station, transmitting certain signals for this purpose. We determine visually headings p to the working radio station related to the longitudinal axis of ship and simultaneously take readings on the radio direction finder, i.e., determine angles q. Knowing p and q, we calculate deviation f and construct the curve of deviation in the form of the dependence of f on q.

The curve of deviation of a ship radio direction finder can be found by the following methods.

1. We use work of the sender of a non-directional radio beacon or auxiliary ship. The ship with the radio direction finder turns every $10\text{-}15^{\circ}$ near the beacon or auxiliary-ship. On every course there is determined visually the heading to the transmitter p and there is made a reading on the radio direction finder q. Deviation is defined as the difference between these readings.

Instead of lie on courses every 10-15°, it is possible to accomplish continuous circulation and to take visual readings and radio direction finder readings every 10-15°.

The distance between the radio direction finder and the sender is more than 2-3 λ, in order to be in the field of radiation of the sender.

2. The ship with the radio direction finder can be turned with the help of an auxiliary tug. Such a method is employed for large ships, whose own movement is costly.

3. The ship with the radio direction finder can stand at anchor, and the auxiliary ship with radio transmitter passes around it. Reading every 10-15° of movement of the auxiliary ship simultaneously visual headings p and radio bearings q, we calculate deviation f.

4. It is possible to take deviation by an invisible radio station. We calculate the bearing α from the point of taking the deviation to the radio station. To determine the true bearing every 10-15° of turn of the ship we determine radio bearings q and simultaneously compass courses (KK). Knowing deviation of the compass ΔK and declination ΔM, we calculate p and f:

$$p=\alpha-(KK+\Delta M+\Delta K) \quad , \quad =p-q.$$

When taking the curve of deviation it is necessary that at a radius of at least one nautical mile there are no other ships or harbour structures.

The power of radio station from which we take the deviation should be sufficient so that readings are absolutely clear. With large range of waves of the direction finder we take deviation on several waves.

Deviation varies with change of the drought of the ship.

So that there is no error from parallax, the distance from the loop of the radio direction finder to instrument, on which we read the heading, should be not more than 1/192 of the distance between the direction finder and the transmitter.

If in the radio direction finder there is no means of compensation for deviation, after determination of deviation and construction of the curve of deviation we calculate coefficients of the expansion of the curve in a Fourier series.

Calculation of coefficients of deviation is reduced to replacement of integral expressions of coefficients of the Fourier series by the sum of ordinates of the curve of deviation at definite intervals of degrees. In Table 10.1 (see insert at the end of the book) there is given the form for calculating coefficients of deviation using ordinates every 15°.

After determining coefficients of deviation we inspect around the loop of the direction finder to see if they are removable factors, causing any abnormally great coefficient (see § 5.8). As a rule, after removal of factors causing deviation it is necessary to again take the curve of deviation.

On the basis of the obtained results for subsequent use we construct curves of deviation or compose tables of deviation for approximately every 5° change of q.

If in the radio direction finder there is equipment to compensate for deviation, then we first determine deviation only on eight headings: 0, 45, 90, ..., 315° at that frequency, at which compensation is provided. We calculate coefficients A, D, and E, which compensate. Then we take residual curves of deviation by one of described methods.

10.2. Radio Direction Finder on an Aircraft

On an aircraft radio direction finder besides normal requirements there is presented an additional one — strength of construction (loop, receiver, etc.) with minimum volume and weight. Such a requirement is set because equipment on an aircraft experiences strong shocks. For combat shocks the receiver is fastened to special shock absorbers.

Aircraft interferences, in two forms — acoustic and electrical — adversely affect actual sensitivity of a radio direction finder.

In light of the impairment of the actual sensitivity from acoustic interferences during sound direction finding, on aircraft they apply visual radio direction finders, most frequently radio compasses and radio compasses.

Basic electrical interferences are interference from ignition and interference from electric generators and motors of the aircraft. The most effective method of counteracting interferences of ignition is complete shielding of the ignition circuit. Other methods do not lead to complete freeing from interferences in the wide band width of the direction finder. Interferences from electrical generators and motors, mainly commutator noises, are cancelled by blocking their circuits with chokes and capacitances, and measures are also taken to decrease sparking.

In spite of all the measures against interferences, the level of noises on aircraft is greater than on earth, and, consequently, sensitivity of a radio direction finder is worse.

The aircraft radio direction finder is often used for flight on a radio station. In this case the rotatable loop is installed across the airplane fuselage (the

goniometer is set at 0°), and the course of aircraft is regulated in such a manner that the loop remained in the position of bearing (in telephones, zero audibility, or on the indicator of the semicompass, a zero reading).

With such use of a radio direction finder from wind there is created drift of the aircraft and time of flight is extended. By selection of a leading course, at which the resultant speed of the aircraft and wind coincide with the direction of flight, it is possible to achieve a direct line of flight; for this we rotate the loop a bit. Correct rotation of the loop can be judged by preservation of the compass reading, if we maintain the course from readings of the radio direction finder. With increase of flight speed the influence of wind drift decreases.

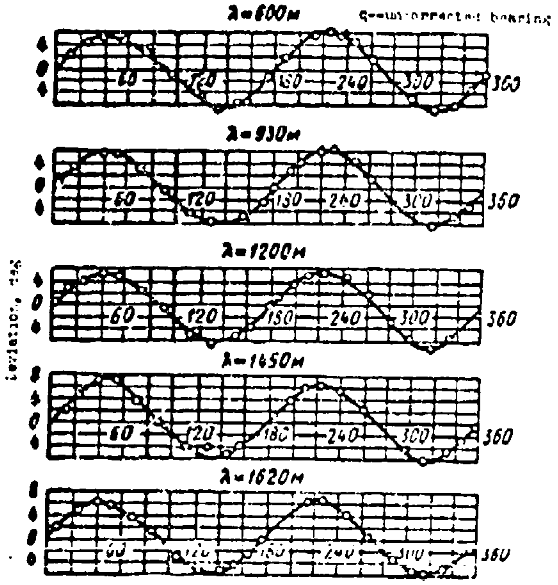

Fig. 10.1. Sample deviation curves.

Orientation and check-out of the mounting of the antenna array of the radio direction finder on an aircraft is carried out just as on a ship. Taking of the deviation of an aircraft radio direction finder in air presents difficulties. Usually we place the aircraft in flight position on the ground on a rotating circle and turn it at different angles to a local transmitter, located a distance of at least 2-3 λ from the aircraft [10.1]. By comparison of visual readings and radio bearings we determine deviation. The site where we take the deviation should be free of disturbing objects (wires, antennas, trees, etc.). The procedure for taking the deviation curve and analysis of results for an aircraft radio direction finder do not differ from the procedure and analysis for a ship radio direction finder.

Sample deviation curves of a medium-wave radio direction finder on an aircraft are given in Fig. 10.1.

10.3. Compensation of Deviation in a Radio Direction Finder with a Rotatable Loop

There exist several methods of compensation of deviation of a radio direction finder. They can be divided into two groups: methods of mechanical compensation of

deviation and; methods of electrical compensation of deviation.

In some installations both methods are used simultaneously.

Mechanical Methods of Compensating Deviation

Principle of work of a mechanical compensator is based on the fact that the dial (or dial indicator) on which the bearing is read fitted on the shaft (of the loop or goniometer) by an auxiliary device. This device creates displacement of the dial (or dial indicator) with respect to the loop or searcher coil of the goniometer an angle, equal to the deviation, and thus deviation is compensated.

In Fig. 10.2 is depicted a system of four levers a, b, c, and d, connected at points 1, 2, 3 by hinges. Between points 1 and 3 there acts a spring, thanks to which hinge 2 and the roller attached to it constantly press on disk L. Lever a is connected to the shaft of the loop (or searcher coil of the goniometer); lever b is connected to the dial indicator. With rotation of the loop shaft the dial indicator tracks the loop.

If disk L has the form of a circle, then the angle between the loop and the dial indicator remains constant, since their relative position does not change. Giving disk L the corresponding form, differing from a circle, it is possible so to move the dial indicator with respect to the loop (leading with an indentation and lagging with a projection) as to compensate any deviation of the radio direction finder.

Sometimes instead of disk L we employ flexible steel tape, whose form is regulated by levers. The roller goes inside tape. The spring draws points 1 and 3 together.

The mechanical compensator can compensate deviation of any law. However, it is advisable to compensate only even coefficients of deviation D, E and constant deviation A. Actually, if one were to compensate semicircular deviation (B and C), then, reading the bearing not from the correct side ($q \pm 180°$), we make an error, equal to

$$2(B \sin q + C \cos q),$$

and with noncompensated semicircle of deviation, if we do not consider deviation, error is equal only to

$$B \sin q + C \cos q.$$

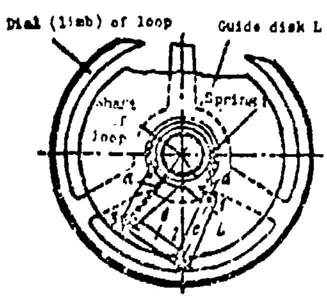

Fig. 10.2. Mechanical compensator of deviation.

The mechanical compensator cancels deviation on one wave. For other waves there are given tables and curves of residual deviation.

Electrical Compensation of Deviation by Installing a Loop

As was shown, of greatest importance in deviation of ship and aircraft radio direction finders, working on medium and long waves, is quadrant deviation. Therefore in practice we are limited to electrical compensation only of this deviation, although in principle we can compensate electrically for other components of deviation. We already saw that the body of a metallic ship (aircraft, dirigible) creates deviation with the same law, as a re-emitting loop. On the basis of this it is possible to compensate deviation caused by the metallic body by installation of an auxiliary frame.

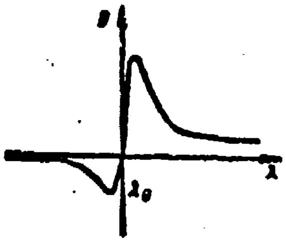

Fig. 10.3. Change of coefficient of deviation D with change of wavelength.

The body of a metallic ship (aircraft, dirigible) is equivalent to a closed loop with characteristic wavelength λ_0, approximately equal to twice the length of the body. The law of change of D with change of the received wave λ is shown in Fig. 10.3. Ideal compensation on all waves could have been achieved by construction around the direction finder loop of a closed loop with parameters, equal to parameters of the frame which is equivalent to the body. In practice, this is unrealizable, since we obtain very large dimensions of the compensating loop. It is possible to construct compensating loops of smaller dimensions, and couple them to the direction finder loop in such a manner as to compensate deviation, e.g., on long waves. Since the fixed wave will be much larger the characteristic wave of such an additional loop, the compensated deviation will almost not vary with change of the waves of direction finding. This will occur until the wave approaches the characteristic wave of the frame, equal to its perimeter. It turns out that by an additional loop, practically relizable, it is possible to compensate coefficient D, constant on the wave range.

For a great many aircraft and ships in their working range of direction finding coefficient D remains constant, and, consequently, for compensation it is possible to choose a compensating closed circuit.

According to formula (5.43) we had

$$D = \frac{m}{2+m},$$

where

$$\dot{m} = \frac{j\omega M_m h_{ok}}{Z_{ok} h_p};\ h_{ok} = \frac{2\pi S_2}{\lambda};\ h_p = \frac{2\pi N S_1}{\lambda};$$
$$Z_{ok} \approx j\omega L_{ok}.$$

Given the area of compensating loop S_2, it is possible to calculate L_{01} and M_{max} and then m (N and S_1 — number of turns and the area of the direction finder loop — are known). From m we calculate coefficient D, compensated by the additional loop. Thus, by approximation it is possible to find the required compensating loop.

Sometimes for compensation they make the additional loop rigid, consisting of two spaced steel rings, fixed motionlessly, symmetrically to the shaft of the loop on both sides of it, along the longitudinal axis of the ship (aircraft).

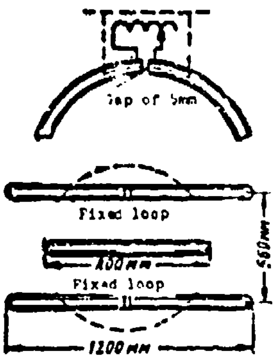

Fig. 10.4. Fixed compensating loops.

By changing the inductance of coil, which closes the rings (Fig. 10.4), it is possible to achieve compensation of the required value of coefficient D. The adjustable magnitude in this construction is Z_{01}.

10.4. Electrical Compensation of Deviation in a Goniometric Radio Direction Finder

Compensation of Quadrant Deviation D sin 2q

Earlier it was shown that if fields created by field coils of the goniometer are not identical and are equal to

$$H_{1\,\text{макс}} \text{ and } H_{2\,\text{макс}} = aH_{1\,\text{макс}},$$

where a ≠ 1, then in the direction finder there appears error, which is expressed (4.16):

$$\operatorname{tg}\Delta = \frac{\frac{a-1}{a+1}\sin 2\theta}{1 - \frac{a-1}{a+1}\cos 2\theta}.$$

The dependence (4.16) of error on the arrival direction of the wave is analogous to quadrant deviation from the hull of a ship (5.57). This view can be used for compensation of quadrant deviation in goniometric radio direction finders, creating in the goniometer an error the opposite of the deviation. Field coils of the goniometer are made identical; to produce of unequal fluxes one of the field coils (longitudinal) is shunted by an inductance of a combine inductance and capacitance.

only an inductance, in which there is created an error opposite to quadrant deviation D sin 2q, and constant in the wave range.

Let us designate:

L_{long} — inductance of longitudinal loop;

L_{trans} — inductance of transverse loop;

L_1 — inductance of each field coil (we assume they are identical);

L_2 — shunting inductance;

h_c — virtual height of each loop.

Then the maximum flux in the goniometer from the transverse loop is equal to

$$H_{1\,\text{макс}}=\frac{kEh_c}{\omega(L_п+L_1)};$$

maximum flux from the longitudinal loop is equal to

$$H_{2\,\text{макс}}=\frac{kEh_c}{\omega\left(L_{пр}+\frac{L_1L_2}{L_1+L_2}\right)}\,\frac{L_2}{L_1+L_2}=aH_{1\,\text{макс}},$$

whence

$$a=\frac{1+\frac{L_1}{L_п}}{\frac{L_{пр}L_1}{L_2L_п}+\frac{L_{пр}}{L_п}+\frac{L_1}{L_п}}. \tag{10.1}$$

If $L_{trans} = L_{long} = L$, then

$$a=\frac{1+\frac{L_1}{L}}{1+\frac{L_1}{L_2}+\frac{L_1}{L}}. \tag{10.1'}$$

Knowing quadrant deviation of the radio direction finder, it is possible to select a in such a manner that

$$a=\frac{1-D}{1+D}, \tag{10.2}$$

i.e., so that there is created a deviation, opposite in sign to deviation from the hull. Then quadrant deviation of radio direction finder will be cancelled:

$$D=\frac{LL_1}{LL_1+2(LL_2+L_1L_2)}. \tag{10.3}$$

From formula (10.1) the expression for shunting inductance L_2 will be

$$L_2=\frac{aL_1L_{пр}}{L_п+L_1(1-a)-aL_{пр}}. \tag{10.4}$$

Usually the coil for shunting longitudinal loop (compensating choke) is made with leads, where they are chosen in such a manner as to compensate the coefficient of quadrant deviation D approximately each degree. Then after taking the deviation

curve it is simple to connect the corresponding lead of the coil.

Since from equations (10.3) and (10.4) it is clear that L_2 does not depend on wavelength, then the compensated quadrant deviation D will not depend on the wave.

Quadrant deviation, not depending on wavelength, is observed with waves, larger than 5-10 lengths of the ship hull.

In this calculation we disregarded capacitance of feeders, connecting the loops to field coils of the goniometer. If we consider these capacitances, then it turns out that the compensated coefficient of quadrant deviation D does not remain constant on different waves, but grows with decrease of wavelength. Sharpness of change of the compensated coefficient with the wave depends on the magnitude of the capacitance of the feeders. In practice the coefficient of quadrant deviation of a ship (aircraft) radio direction finder also grows with decrease of the wavelength. On long waves the capacitance of feeders is small, and the compensating choke can be calculated by formula (10.4).

Sometimes after connection of the compensational choke designed for the longest wave and determination on several waves of residual deviation it is found that on shorter waves there is a coefficient of quadrant deviation, increasing with decrease of wavelength. The means that abruptness of change of the compensated coefficient D with the wave is less than abruptness of change with the wave of real coefficient D. In such a case, in order to achieve the best compensation of quadrant deviation in the wave range, it is necessary to couple an additional capacitance in the transverse loop. Conversely, if steepness of variation of the compensated coefficient D with the wave is greater than steepness of variation of the real coefficient D, then one should couple the capacitance in the longitudinal loop [10.2].

In a two-path automatic radio direction finder it is possible to compensate the coefficient of quadrant deviation D, constant in the frequency range, by employing unequal gain factors in the channels.

Compensation of Quadrant Deviation E cos 2 q

During the analysis of the goniometric system we saw that presence of coupling between field coils of the goniometer leads to the appearance of quadrant error of form E cos 2q (§ 4.6), where (4.27)

$$E = \left(-\frac{Z_0}{Z}\right) \text{ radians,}$$

where Z_c — coupling impedance between field coils;

Z — impedance of loop circuit;

$\frac{Z_c}{Z}$ — has totally real value, if we disregard active components of Z_c and Z.

Simultaneously there appears octant error $K \sin 4q$, where

$$K = -\frac{1}{2}\left(\frac{Z_c}{Z}\right)^2 \text{ radians.}$$

By choice of coupling impedance Z_c it is possible to compensate quadrant deviation $\mathbf{E} \cos 2q$.

The simplest method of creating of coupling between field coils is connecting inductances (compensational chokes) between the field coils (Fig. 10.5).

In Fig. 10.5a we designate:

L — inductance of loop;

L_1 — inductance of field coil;

L_2 — inductance of coupling coil (compensational choke);

L_3 — inductance of searcher coil.

Let us disregard active resistances of loop circuits. We determine the expressions for Z and Z_c. We designate (Fig. 10.5b) resistance of the part of the circuit below points aa, but above points bb by X. Impedance of the loop circuit is

$$Z = j(\omega L + X).$$

We find the expression for Z_c from the influence of circuit II on circuit I:

$$Z_c = j(\omega L + X)\frac{LL_1}{LL_1 + 2(LL_2 + L_1L_2)}.$$

Compensated coefficient of deviation will be

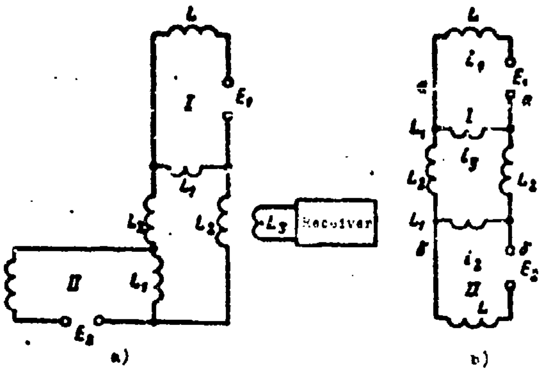

Fig. 10.5. Compensation of deviation E cos 2q.

$$\mathbf{E} = \frac{Z_c}{Z} = \frac{Z_c}{j(\omega L + X)} = \frac{LL_1}{LL_1 + 2(LL_2 + L_1L_2)}. \qquad (10.5)$$

Formulas (10.4) and (10.5) for compensated coefficients of deviation **E** and **D** are absolutely identical.

Consequently, inductance of coils L_2 for compensation of deviation $\mathbf{E} \cos 2q$ should be calculated just as inductance of the coil for compensation of deviation $\mathbf{D} \sin 2q$.

A ship radio direction finder with a goniometer system usually is equipped with three identical compensational chokes with leads for compensation of deviation every 1°. One choke serves to compensate coefficient of deviation **D**, and we connect it in parallel to the longitudinal loop of the radio direction finder. Two other chokes serve to compensate coefficient of deviation **E**; they are connected between terminals of the field coils.

In comparison of mechanical and electrical methods of compensation of deviation one should note the basic deficiency of mechanical systems — lowering of accuracy of reading of bearing.

Let us assume that the design and electrical circuit of a radio direction finder permit with a given certain field strength of the fixed radio station, wavelength and average observer a certain accuracy of bearing reading (error Δq).

By a mechanical deviation compensator we compensate quadrant deviation

$$f = \mathrm{D}\sin 2q.$$

The true bearing, corrected by the mechanical compensator, is

$$p = q + f = q + \mathrm{D}\sin 2q.$$

Error of reading Δp will be obtained by differentiation of p with respect to q:

$$\Delta p = \Delta q\,(1 + 2\mathrm{D}\cos 2q).$$

When q = 45, 135, 225 and 315°, $\Delta p = \Delta q$, i.e., accuracy of reading does not change.

When q = 0, 90, 180 and 270°, $\Delta p = \Delta q(1 \pm 2\mathrm{D})$, where **D** is expressed in radians.

Accuracy of the taken bearing depends on **D**. Thus, if $\mathrm{D} = 15^\circ$; $2\mathrm{D} = 30^\circ = \frac{\pi}{6} =$ $= 0.523$ rdn, then $\Delta p = 1.5\Delta q$.

With growth of **D** inaccuracy of reading of bearing on a radio direction finder with a mechanical compensator of deviation increases, so that work will be especially bad when **D** is great.

Electrical methods of compensation do not possess this deficiency.

The mechanical compensator of deviation creates additional error when there is an angle of silence (oscillation of the pointer of the indicator because of noises, etc.). Indeed let us assume that on the dial of the radio direction finder limits of the angle of silence will be

$$p_1 = q_1 + \mathrm{D}\sin 2q_1,$$
$$p_2 = q_2 + \mathrm{D}\sin 2q_2.$$

True bearing is equal to

$$p = \frac{q_1 + q_2}{2} + \mathrm{D}\sin(q_1 + q_2).$$

Bearing on dial p_0 is defined as the arithmetic mean of p_1 and p_2:

$$p_0 = \frac{p_1 + p_2}{2} = \frac{q_1 + q_2}{2} + \frac{D}{2}(\sin 2q_1 + \sin 2q_2)$$

or

$$p_0 = \frac{q_1 + q_2}{2} + D\sin(q_1 + q_2)\cos(q_1 - q_2).$$

Error in reading of bearing will be determined from expression

$$|\Delta| = |p_0 - p| = D\sin(q_1 + q_2)[1 - \cos(q_1 - q_2)].$$

For any angle of silence, determined by $q_1 - q_2$, maximum error will be at $q_1 + q_3 = 90°$. Then

$$|\Delta_{\text{макс}}| = D[1 - \cos(q_1 - q_2)] = 2D\sin^2\left(\frac{q_1 - q_2}{2}\right).$$

If $D = 20°$, $q_1 = 65°$, $q_2 = 25°$, i.e., angle of silence is 40°, then

$$|\Delta_{\text{макс}}| = 40° \sin^2 20° = 4°,7.$$

10.5. Land (Airport, Shore) Radio Direction Finder

The antenna system of a ground radio direction finder should be installed as far as possible from re-emitting, current-conducting objects on a site with good ground conductivity. At the place of installation of a radio direction finder there should be no industrial interferences, impairing its sensitivity and accuracy. The place should be selected so that it is convenient as far as access roads, power supply, communication circuits, and so forth. It should be suitable for work in the RDF network (see Chapter 11).

Let us discuss in greater detail the fitness of the place from the point of view of ground conductivity and surroundings.

Requirements for ground conductivity are determined by the type of antenna system of the radiodirection finder. This question is considered in Chapter 6. If there is installed an antenna system, whose feeders must be protected from reception of a horizontally polarized electrical field, for instance a U-shaped goniometric or phase system of spaced vertical antennas, the ground conductivity on the antenna site should be at least $10^{-2}\,\frac{1}{\text{ohm}\cdot\text{m}}$ if the feeders are buried in the ground, and at least $10^{-3}\,\frac{1}{\text{ohm}\cdot\text{m}}$ if there is applied a ground metallizing network.

Ground conductivity can vary from month to month. For determination of average or minimum ground conductivity by separate measurements see [10.8]. Ground conductivity should be measured not only on the surface, but at the depth of laying

of feeders (about 2 m with burying of feeders). During measurements it is necessary to consider the depth of penetration of electromagnetic waves in the soil. In addition to Fig. 6.5 in Table 10.2 there are given for soils two values of depth conductivity where field strength decreases to 1/10 with respect to the field at the surface. One should also turn attention to whether ground conductivity is identical within the antenna site (see § 5.2).

In Table 10.3 there are given permissible minimum distances to certain reverse emitters or permissible angles of visibility when using an antenna system with a cosine directivity pattern [10.6]. With increase of separation of antennas errors from the influence of reverse emitters decrease (§ 5.3) and, correspondingly, distance to emitters can be decreased. Distances of Table 10.3 are given separately for long and medium waves (range of frequencies 100 Kc to 1.5 Mc), short waves (range of frequencies 1.5 to 30 Mc) and ultrashort waves (30 to 400 Mc) and for two cases:

Table 10.2. Depth of Penetration of Radio Waves, at Which Field Strength Decreases to 1/10.

Frequency, Mc	Depth of penetration, m	
	Bad conductivity $10^{-3} \frac{1}{ohm \cdot m} = 10^7$ CGSE	Normal conductivity $10^{-2} \frac{1}{ohm \cdot m} = 10^8$ CGSE
0.1	48	15
0.3	29	9
0.5	23	7
1.5	15	4
10	11	2.1
30	11	1.8
300	11	1.8

a) when expected mean quadratic error σ from the influence of one reverse emitter σ has a value on long, medium and short waves of 1°, and in UHF range 0.5°;

b) when expected mean quadratic error σ from influence of reverse emitter has value on long, average and short waves of 5°, and in UHF range 2°. The easier requirements pertain to the case when it is possible to permit the worst accuracy in light of other advantages of the place.

Simultaneously with errors there can be observed blurred minima during sound direction finding and an elliptical image of bearing in a two-channel visual radio direction finder, if phases of fields of the forward and reverse emitters do not coincide. In composing Table 10.3 it was assumed that the re-emitter creates maximum possible error without impairing the actual reading.

The table was composed on the basis of experimental materials and calculating formulas of errors from reverse emitters, having regular geometric forms (sphere, hemisphere, cube, mirror surface, vertical wire, horizontal wire, etc.).

Table 10.3. Requirement to Ensure Normal Work of a Radio Direction Finder

Number of point	Causes of error and parameters limited by them	Admissible values: Average, medium waves (100 Kc to 1.5 Mc)		Short waves (1.5 to 30 Mc)		Ultrashort waves (30 to 400 Mc)	
		$\sigma = 1^{\circ}$	$\sigma = 5^{\circ}$	$\sigma = 1^{\circ}$	$\sigma = 5^{\circ}$	$\sigma = 0.5^{\circ}$	$\sigma = 2^{\circ}$
I	Slope of section, deg	0.5	2	0.5	2	0.5	1
II	Vertical conductors — for a grounded one of length $l = 0.1\lambda$ (vertical angle, deg)	10					
	— for grounded one $l = 0.25\lambda$ — for ungrounded one $l = 0.5\lambda$ (distance)	$7l$	l	$7l$	$2l$	$15l$	$4l$
	— for ungrounded one length several λ (distance)			15 m		30 m	
III	Slanted conductors — in general (distance)	Approximately as in point II					
	— angle of inclination 45° (distance)	Half as large as in point II					
IV	Horizontal wires — telegraph at height h = 6-7 m (distance, m)	180-200	45	See medium wavelengths			90
	— higher suspension H (distance, m)	$(180-200)\frac{H}{h}$	$45\frac{H}{h}$				
V	Wire barrier (distance, m)	Does not influence if grounded		90	45	180, 30 [10.7]	90
VI	Large conducting objects — square in area $\lambda \times \lambda$ (distance, λ)	See point IX		25	5	50	15
	— other area (distance, λ)	Varies in proportion to area of object					

(Table 10.3 continued)

Number of point	Causes of error and parameters limited by them	Admissible values					
		Average, medium waves (100 Kc to 1.[illegible] Mc)		Short waves (1.5 to 30 Mc)		Ultrashort waves (30 to [illegible])	
		$\sigma = 1^{\circ}$	$\sigma = 5^{\circ}$	$\sigma = 1^{\circ}$	$\sigma = 5^{\circ}$	[illegible]	$\sigma = 2^{0}$
VII	Forest —single trees —large group of trees (distance, m)	45 350	20 90	90 700	20 90	90 360	45 180
VIII	Isolated hills (vertical angle, deg)	2	5	2	5	Should not black out	
	Mountainous site	Not fit for installation of RDF					
IX	Buildings —small nonconducting (vertical angle, distance) —large conducting ones (distance, m)	2°	90 m 400	800 m			
X	Buildings housing RDF	If antennas are mounted in building it should be wooden					
XI	Takeoff and landing reinforced concrete airport runways (distance)	45 m	No influence	λ	No influence	λ	No influence
XII	River (distance, m)	No influence		No influence		300	No influence
XIII	Ditches, embankments (distance, m)	30	30	30	30	90	30
XIV	Railroad rails (distance, m)		90		90		

If the distance to the reverse emitter differs from that shown in the table, one should consider that field strength from the reverse emitter, and consequently also the error vary on long, medium and short waves: if the distance to the emitter is less than 0.1λ it is inversely proportional to cube of distance D (by dependence $1/D^3$); if distance is more than λ, it varies as $1/D$; in range of frequencies above 30 Mc with distance of several wavelengths, it varies as $1/D^2$.

Total mean quadratic error from several reverse emitters, each of which causes error σ_1, σ_2, σ_3, ..., is most correctly calculated by the formula

$$\sqrt{\sigma_1^2+\sigma_2^2+\sigma_3^2+\cdots}$$

The site chosen for installation of the antenna system after external inspection at a radius of several wavelengths usually is inspected for faults by a portable radio direction finder and local transmitter, carried or transported on motor vehicle, ship, or aircraft around the antenna system, or by direction finding of radio stations whose locations are known. The site is considered suitable for a radio direction finder, if mean quadratic error does exceed tolerable limits. So that it is possible to use a local transmitter, around its antenna at a radius of about 0.5 λ there should not be reverse emitters.

Distance from the transmitter to the center of the antenna system of the radio direction finder should be at least $(1-1.5)\lambda_{max}$ and 3-4 times greater than the separation of antennas in the antenna system of the radio direction finder.

Inspection of the site permits determining the order of errors expected during operation.

When constructing the home for a direction finder one should avoid construction of metal drain pipes and lightning rods. At a distance of about 150 m from the house electrical and telephone wiring must go into underground cable.

If the antenna system of vertical antennas is mounted around the location with the receiver-indicator, then power and communication cables should be brought in symmetrically to a pair of adjacent feeders, on the bisector of the angle between them.

Wiring inside the building should be very close to the floor, avoiding creation of loops (frames). If the radio direction finder is installed together with the transmitter, the antenna of the latter should be mounted symmetrically to the outdoor array of the direction finder, whose vertical antennas are set around the building. In the case of installation of several outdoor arrays on one site to solve the problem of the permissible minimum distance between antenna systems one should be guided by the fact that zenith angle at which one can see outdoor display from the center of the other, should be no more than 2-3°.

With a slanted site with slope permissible according to Table 10.3 vertical antennas in the system should be set normally to the plane of the site.

In a hilly site the best place is summit of a secluded round hill, which dominates the remaining hills.

The installed radio direction finder before use is calibrated by a local transmitter, observing here the shown requisite distances between transmitter and antenna system and satisfying requirements on surroundings of the transmitter antenna, or by distant radio stations.

The radio direction finder is calibrated also periodically during operation to check its accuracy. To permit periodic calibration dimensions of the direction finder site which are free from re-emitters are increased, since aound the antenna of the transmitter there also should be no reverse emitters. As a result of analysis of calibration materials we obrain an estimator of the accuracy of direction finding. If the obtained constant errors are confirmed by operation specifications of the radio direction finder, they are considered corrections to bearings. Sometimes results of calibration help us to find and remove the cause of large errors.

It is necessary to consider that errors from a local transmitter can differ from errors from distant radio stations because of noncoincidences of relative phases of fields of forward and reverse radiation when fixing local and distant transmitters.

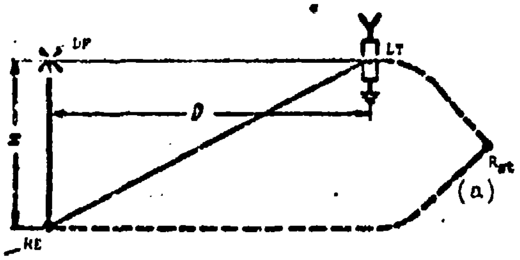

Fig. 10.6. Direction finding of near and distant transmitters. KEY: (a) [Designation uncertain, R_{st} could be radio station].

Let us determine such a minimum distance to a local transmitter that errors from local and distant transmitters differ by not more than 10% (Fig. 10.6).

In the figure we designated:

DF — radio direction finder;

RE — reverse emitter at distance X from direction finder;

LT — local transmitter at distance D from direction finder (D >> X).

Difference of phases of fields of forward and reverse radiator during direction finding of a distant radio station will be

$$\varphi_1=\frac{2\pi X}{\lambda}.$$

The same difference of phases from a local transmitter is

$$\varphi_2=\varphi_1+\Delta\varphi=\frac{2\pi}{\lambda}[X+(\sqrt{D^2+X^2}-D)]\approx\frac{2\pi}{\lambda}\left(X+\frac{X^2}{2D}\right)$$

or

$$\Delta\varphi\approx\frac{\pi X^2}{\lambda D}.$$

Error in direction finding is proportional to the cosine of the difference of phases of the fields of the forward and reverse radiators.

To satisfy the formulated requirement for permissible difference of errors during direction finding of local and distant transmitters, condition $\frac{\cos\varphi_2}{\cos\varphi_1} \geq 0.9$ should be satisfied. Since, for $\frac{\pi}{2} > \varphi_1 > 0$ $\cos\varphi_2 = \cos(\varphi_1 + \Delta\varphi) \approx \cos\varphi_1 - \Delta\varphi\sin\varphi_1$, $\frac{\cos\varphi_2}{\cos\varphi_1} = 1 - \Delta\varphi\tan\varphi_1$ and $\Delta\varphi\tan\varphi_1 \leq 0.1$ or we should have

$$D \geqslant \frac{10\pi X^2}{\lambda} \operatorname{tg}\varphi_1.$$

Consequently, when $X = 100$ m we should have conditions:

for $\lambda = 300$ m, $D \geq 1.7$ km; for $\lambda = 12$ m, $D \geq 45$ km.

On short waves these requirements are not practically feasible. Even if calibration is performed at the required distance, error varies with the frequency, azimuth and height of location of the transmitter. In Table 10.4 there are calculated changes of frequency, azimuth and angle of inclination of the wave front, corresponding to change of error of bearing from the influence of the reverse radiator from a maximum value down to zero. It is assumed that the reverse radiator is a distance of 100 m from the antenna system of the radio direction finder.

Table 10.4. Changes of Frequency, Azimuth and Angle of Inclination of the Wave Front, Corresponding to Change of Error from a Maximum to Zero

Frequency, Mc	1 Mc	25 Mc
Change of frequency, %	37	1.48
Change of azimuth, deg	44	1.75
Change of angle of inclination of wave front, deg	75	15

From the presented it follows that calibration by a local transmitter cannot reveal systematic errors of the radio direction finder, which are found by selection of bearings by distant radio stations, whose locations are known, and as a result of prolonged study of the operation of the radio direction finder.

MT-65-58
Principles of Radio Direction Finding,
"Soviet Radio" Publishing House,
Moscow, 1964.
Pages: 571-597

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CHAPTER 11

ACCURACY OF POSITION FINDING BY RADIO BEARINGS

List of Designations Appearing in Cyrillic

к = circ = circular

мин = min = minimum

э = ell = ellipse
э = op = operational

П = RDF = radio direction finder

In order to obtain good results in direction finding, it is necessary to have radio direction finders, possessing sufficient accuracy and sensitivity, which should be properly situated and correctly used. It is necessary also to be able for each fix to estimate bearings and to find from bearings and estimations of them the most likely place or region for finding the fixed target.

In the present chapter we consider methods of appraisal of a single bearing and accuracy of position finding from bearings of n radio direction finders, and also methods of construction of working zones of two radio direction finders.

At the base of chapter lies application of the statistical theory of errors to radio direction finding.

11.1. Methods of Estimating a Single Bearing

Errors during direction finding, as shown in § 2.4, are divided in errors of the system, which can be allowed for in the form of corrections, and random, which cannot be allowed for by corrections. Random errors characterize an individual reading of bearing.

In order to determine the position of a target and to calculate position error from bearings of n radio direction finders, it is necessary to know estimates of

errors of bearings. On the basis of large number of observations, conducted over a prolonged time, there can be found the mean quadratic operational error of a radio direction finder σ_{op}. However, estimated mean quadratic error of an individual bearing may differ considerably from σ_{op}, especially on short waves.

Sometimes we use subjective estimation of the bearing by the operator, for which we beforehand establish 4-5 categories of bearings (for instance, by quality of readings, by degree of their stability, etc). For each category of bearing we experimentally establish the mean quadratic angular error. Subjective estimation with known accuracy can be applied only on ultrashort, medium, and long waves, since on these wave ranges mean quadratic error for the outlined categories of bearings should be preserved. On shortwaves, where there are too many factors affecting the accuracy of direction finding, subjective estimation may lead to incorrect results.

It is better to employ objective estimation of bearings [11.4, 11.5], which is based on the physical picture of propagation and conditions of direction finding of short waves. We give in [11.5] a method of estimating bearings on short waves. Analysis of random errors of a short-wave radio direction finder shows that they can be divided into three statistically non-connected groups with dispersions V_1, V_2, V_3, deg^2.

1. Errors varying very slowly in time — error of instruments and from the influence of the position (near and distant surroundings). For the dispersion of these errors it is possible to write:

$$V_1 = V_{11} + V_{12}(f),$$

where V_{11} — a component which does not depend on frequency. V_{11} varies from 0.1 deg^2 to 1 deg^2 depending upon the antenna system of the radio direction finder and the site of its installation; $V_{12}(f)$ — component, depending on frequency. On the basis of operational specifications one may assume that $V_{12}(f) \approx Ab$, where A — constant coefficient, depending on the quality of the installation site and separation of the antenna system; b — coefficient, dependent on frequency:

- for frequencies 2-4 Mc, $b \approx 1$;
- for frequencies 4-9 Mc, $b \approx 2$;
- for frequencies above 9 Mc, $b \approx 3$.

2. Errors varying slowly in time — errors from lateral deviation of radio waves during reflection from the ionosphere. This error depends on the distance and, to an extent, on the time of day; because of the large period of change this error usually is not averaged during the time of taking a bearing (§ 6.4). Values

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of dispersion of lateral deviation V_2 deg^2 = $\Phi(D)$ are given in Fig. 6.18 for a small-base antenna system.

3. Errors varying rapidly in time — errors from interference of radio waves and polarization; here there enters subjective error of operator readings. These errors are averaged by the operator. The operator usually takes several (5-12) averaged readings, each of which is the result of observations for 5-10 sec. With greater duration of each separate averaged reading the operator, who, as a rule, recalls the picture of the bearings only for the last 5-10 sec, will waste part of the time of direction finding.

Let us assume that the operator took n averaged readings, obtained average bearing θ, the difference between maximum and minimum values of averaged reading r deg, and observed during the time of taking of a separate averaged reading variation of the bearing δ deg.

Dispersion V_3 is calculated by the formula

$$V_3 = V_{31} + V_{32}.$$

Proceeding from the normal law of distribution of errors of bearings

$$V_{31} \approx \frac{r^2}{n}, \ \text{deg}^2.$$

Component of dispersion V_{32} considers subjective error of the operator. It depends on limits of variation of bearing δ during the time of taking averaged readings on a cathode-ray tube or on the angle of silence during a sound method of direction finding by a minimum.

In [11.5] it is proposed to determine V_{32} on the basis of magnitude of play of the bearing during the time of an averaged reading or the angle of silence within limits:

$$\begin{array}{rl}
0\text{—}3^\circ & V_{32} = 0 \ \text{deg}^2; \\
9\text{—}13^\circ & V_{32} = 1 \ \text{deg}^2; \\
14\text{—}18^\circ & V_{32} = 2 \ \text{deg}^2; \\
19\text{—}23^\circ & V_{32} = 4 \ \text{deg}^2; \\
24\text{—}37^\circ & V_{32} = 6 \ \text{deg}^2; \\
\text{to above } 38^\circ & V_{32} = 9 \ \text{deg}^2;
\end{array}$$

Thus, total dispersion of error of bearing will be

$$V = V_1 + V_2 + V_3,$$

where

$$V_1 = V_{11} + V_{12}(f), \ V_2 = \Phi(D), \ V_3 = V_{31} + V_{32}.$$

Dispersion V, or mean quadratic error $\sigma = \sqrt{V}$, together with mean bearing θ are used for plotting.

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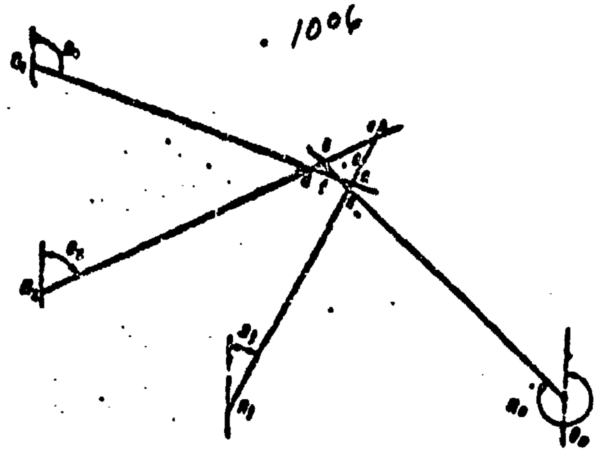

Fig. 11.1. Plotting bearings.

The indicated coefficients for calculation of components of dispersion are empirical and must be definitized on the basis of a check direction finding of radio stations, whose positions are known.

11.2. Ellipse of Error with n Radio Direction Finders

Let us assume that a radio station, located at point O (Fig. 11.1) is fixed by n radio direction finders; $(RDF)_1$, $(RDF)_2, \ldots$, $(RDF)_n$, and there are obtained bearings θ_1, θ_2, ..., θ_n. From angular errors of bearings as a result of plotting we formed a polygon of fixing (abcdef).*

We designate:

Δ_1, Δ_2, ..., Δ_n — angular errors of radio bearings;

σ_1, σ_2, ..., σ_n — mean quadratic angular errors of bearings.

In Fig. 11.2 there is depicted line of bearing of one (j-th) radio direction finder. Let us place the origin of coordinates at point O, the true position of the radio station, axis OX we direct along parallel, axis OY along the meridian of point O. Let us designate by p_j the length of the perpendicular from point O to the line of bearing $(RDF)_jK$: $p_j = D_j\Delta_j$, where D_j — distance between $(RDF)_j$ and O.

Let us assume that as a result of plotting lines of bearings on a map for the position of the radio station we take point S with coordinates x, y, at distance q_j from line of bearing $(RDF)_jK$.

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*We consider that bearings can be plotted in the form of straight lines.

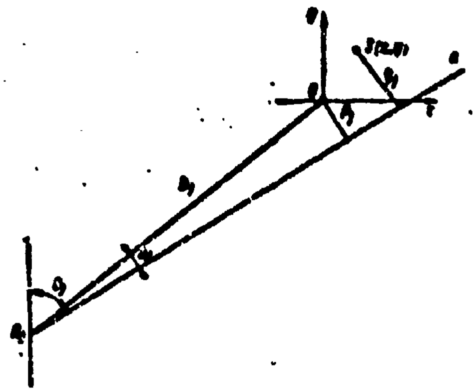

Fig. 11.2. Calculating probability density.

For q_j it is possible to write:

$$q_j = p_j + x\cos\theta_j - y\sin\theta_j.$$

The probability that the distance from point S to line of bearing $(RDF)_j K$ lies between q_j and $q_j + dq_j$, is determined by formula

$$P(q_j)\,dq_j = \frac{1}{\sqrt{2\pi}\,\sigma_j D_j}\, e^{-\frac{q_j^2}{2\sigma_j^2 D_j^2}}\, dq_j,$$

where $\sigma_j D_j = E_j$ – mean quadratic deviation of the line of bearing from the true position of the radio station.

Analogous expression can be written for other lines of bearings. Total probability that the assumed location of the radio station S is found from n lines of bearings, taken from points $(RDF)_1$, $(RDF)_2$, ..., $(RDF)_n$, at distance from q_1 to $q_1 + dq_1$, from q_2 to $q_2 + dq_2$, ..., from q_n to $q_n + dq_n$, correspondingly, will be

$$P(q_1, q_2, \ldots, q_n)\,dq_1 dq_2 \ldots dq_n = \frac{1}{(2\pi)^{\frac{n}{2}} E_1 E_2 \ldots E_n}\, e^{-\frac{1}{2}\sum_{j=1}^{n}\frac{(p_j + x\cos\theta_j - y\sin\theta_j)^2}{E_j^2}}\, dq_1 dq_2 \ldots dq_n. \qquad (11.1)$$

From the principle of least squares coordinates (x_0, y_0) of the most probable position of the fixed radio station are found from the condition of a maximum of expression (11.1) or a minimum of exponent e. To determine x_0, y_0 one should equate to zero derivatives with respect to x and y of exponent e in (11.1). We obtain two equations:

$$\sum_{j=1}^{n}\frac{p_j\cos\theta_j}{E_j^2} + Ax - By = 0,$$

$$\sum_{j=1}^{n}\frac{p_j\sin\theta_j}{E_j^2} + Bx - Cy = 0,$$

from which it follows that

$$\left.\begin{aligned} x_0 &= \frac{1}{AC - B^2}\sum_{j=1}^{n}\left[p_j\,\frac{B\sin\theta_j - C\cos\theta_j}{E_j^2}\right], \\ y_0 &= \frac{1}{AC - B^2}\sum_{j=1}^{n}\left[p_j\,\frac{A\sin\theta_j - B\cos\theta_j}{E_j^2}\right]. \end{aligned}\right\} \qquad (11.2)$$

where

$$\left.\begin{aligned} A &= \sum_{i=1}^{n} \frac{\cos^2 \theta_i}{E_i^2}; \\ B &= \sum_{i=1}^{n} \frac{\sin \theta_i \cos \theta_i}{E_i^2}; \\ C &= \sum_{i=1}^{n} \frac{\sin^2 \theta_i}{E_i^2}. \end{aligned}\right| \qquad (11.3)$$

Point (x_0, y_0) the most likely place of finding the radio station and is called the center of probability.

Let us transfer the origin of coordinates to point (x_0, y_0). Then, the probability of finding the radio station at any point with coordinates (x, y) will be

$$P(x, y)\,dx\,dy = \frac{\sqrt{AC - B^2}}{2\pi} e^{-\frac{1}{2}(Ax^2 - 2Bxy + Cy^2)}\,dx\,dy. \qquad (11.4)$$

From the right side of equality (11.4) we see that only exponent e depends on variables x and y. Consequently, probability P varies only with change of exponent e. Assuming that the exponent is equal to a constant, we obtain the equation of a contour, on whose boundaries the probability of finding the radio station within limits of elementary areas dxdy has a constant value, i.e., the equation of the contour of constant probability density. Let us designate exponent e by coefficient $-\frac{1}{2}K_0^2$. Then expression

$$Ax^2 - 2Bxy + Cy^2 = K_0^2 \qquad (11.5)$$

determines the locus of points with identical probability density P(x, y) dxdy = = const. Equation (11.5) is the equation of an ellipse with its center at point (x_0, y_0), i.e., contours of constant probability density are ellipses with their center of the center of probability.

To determine the integral probability P_{ell} of finding the object of direction finding inside the ellipse constructed for the given value of K_0, one should integrate the expression for differential probability (11.4) within the area of the ellipse S_{ell}:

$$P_0 = \iint_{S_0} \left[\frac{\sqrt{AC - B^2}}{2\pi} e^{-\frac{1}{2}(Ax^2 - 2Bxy + Cy^2)}\right] dx\,dy.$$

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We replace axes x, y by axes x', y', turning them angle γ, where $\tan 2\gamma = \frac{2B}{C - A}$.

Then we replace axes x', y' by axes x'', y'':

$$x'' = \frac{x'}{a_0}, \quad y'' = \frac{y'}{b_0},$$

where a_0 and b_0 — semiaxes of a unit ellipse.

$$a_0, b_0 = \frac{\sqrt{2}}{\sqrt{A + C \mp \sqrt{(A-C)^2 + 4B^2}}}.$$

Let us turn to polar coordinate axes

$$\rho^2 = x''^2 + y''^2$$

and

$$dx''dy'' = \rho d\rho d\varphi.$$

Making these replacements, we obtain the following expression for integral probability P_{ell}:

$$P_0 = \frac{1}{2\pi}\int_0^{2\pi} d\varphi \int_0^{K_0} e^{-\frac{1}{2}\rho^2} \rho d\rho.$$

We consider $\rho^2 = 2u$, then $\rho d\rho = du$:

$$P_0 = \frac{1}{2\pi}\int_0^{2\pi} d\varphi \int_0^{0.5K_0^2} e^{-u} du.$$

Integral probability is determined by formula

$$P_0 = 1 - e^{-0.5K_0^2},$$

from which

$$K_0 = \sqrt{-2\ln(1 - P_0)}. \qquad (11.6)$$

Thus, K_0 in equation (11.5) is calculated by formula (11.6), depending upon the given integral probability P_{ell}. Below we give coefficients K_0 for various values of P_{ell} (Fig. 11.3 and Table 11.1).

If $K_0^2 = 1$, we obtain the equation of the so-called unit ellipse $Ax^2 - 2Bxy + Cy^2 = 1$, and $P_{ell} = 39.4\%$.

We combine axes of the right-angle system of coordinates with principal axes

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of the ellipse of probability. Equation (11.5) of the ellipse takes the form

$$\frac{x^2}{a_0^2}+\frac{y^2}{b_0^2}=K_s^2,$$

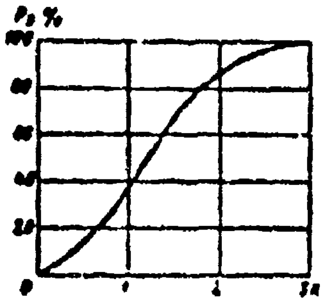


Fig. 11.3. Curve of integral probability.

Table 11.1

P_s	K_s	P_s	K_s	P_s	K_s
0	0	0,3	0,842	0,7	1,552
0,06	0,322	0,394	1	0,865	2
0,1	0,458	0,5	1,179	0,989	3
0,12	0,5	0,6	1,353		
0,2	0,671	0,632	1,41		

where a_0 and b_0 — semimajor and semiminor axes of any ellipse. On the basis of (11.5) expressions for semiaxes of the ellipse will be

$$a_0,\ b_0=\frac{\sqrt{2}K_s}{\sqrt{A+C\mp\sqrt{(A-C)^2+4B^2}}}. \tag{11.7}$$

Upper sign (-) pertains to semiaxis a_0 (major), lower sign (+) pertains to semiaxis b_0 (minor).

We substitute in (11.7) formula (11.6), and after transformations we obtain

$$a_0,\ b_0=\sqrt{-\ln(1-P_s)}\,\frac{\sqrt{A+C\pm\sqrt{(A-C)^2+4B^2}}}{\sqrt{AC-B^2}}. \tag{11.8}$$

Angle γ between the major axis of the ellipse $2a_0$ and the meridian is determined from (11.5) expression

$$\operatorname{tg}2\gamma=\frac{2B}{C-A} \tag{11.9}$$

or

$$\operatorname{tg}\gamma=\frac{A-C-\sqrt{(A-C)^2+4B^2}}{2B}. \tag{11.9'}$$

For two radio direction finders formula (11.8) for calculation of semiaxes of the ellipse of probability is simplified and takes the form

$$a_0,\ b_0=\sqrt{-\ln(1-P_s)}\times$$
$$\times\frac{D_1\sigma_1D_2\sigma_2}{\sqrt{D_1^2\sigma_1^2+D_2^2\sigma_2^2\mp\sqrt{(D_1^2\sigma_1^2+D_2^2\sigma_2^2)^2-4D_1^2\sigma_1^2D_2^2\sigma_2^2\sin^2\alpha_{12}}}}, \tag{11.10}$$

where $\alpha_{12}=(\theta_2-\theta_1)$ — angle of intersection of bearings.

For the angle of orientation of the major axis of the ellipse from (11.9) we obtain

$$\operatorname{tg} 2\gamma = \frac{D_i^2 e_i^2 \sin 2\theta_i + D_j^2 e_j^2 \sin 2\theta_j}{D_i^2 e_i^2 \cos 2\theta_i + D_j^2 e_j^2 \cos 2\theta_j}. \tag{11.11}$$

Knowing a_0 and b_0, we can find the area of the ellipse of probability:

$$S_0 = \pi a_0 b_0 = 2\pi \ln (1 - P_0) \frac{1}{\sqrt{AC - B^2}}. \tag{11.12}$$

For two intersecting bearings the area of the ellipse will be

$$S_0 = \pi \ln (1 - P_0) \frac{E_i E_j}{\sin \sigma_{ij}}. \tag{11.13}$$

Using formulas (11.8) and (11.9'), we can obtain the following expressions for semiaxes of the ellipse:

$$\left.\begin{aligned} a_0^2 &= \frac{K_0^2}{C + B \operatorname{tg} \gamma}, \\ b_0^2 &= \frac{K_0^2}{A - B \operatorname{tg} \gamma}. \end{aligned}\right\} \tag{11.14}$$

In [11.7] there is offered a graphical method of determination parameters of the ellipse of probability using a special plotting board and formulas (11.9) and (11.14) at distances, where bearings can be plotted in the form of straight lines (see Table 12.1).

The method is as follows.

Let us assume that as a result of plotting bearings we have found the center of probability O (Fig. 11.4a). Let us pass to O from the point of location of the radio direction finder $(RDF)_j$ line of bearing $(RDF)_j O$ at angle θ_j to meridian OM. We pass a second line $(RDF)_j K$ at angle σ_j to line $(RDF)_j O$. We place at O a transparent plotting board with rectangular axes, where axis OY is matched with the meridian at O. We mark the points a_j and b_j of the intersection of line $(RDF)_j K$ with axes OY of the plotting board.

From Fig. 11.4a it follows that

$$D_j e_j \approx OL, \quad Oa_j = \frac{D_j e_j}{\sin \theta_j} \quad \text{and} \quad Ob_j = \frac{D_j e_j}{\cos \theta_j}.$$

Axes OY and OX on the plotting board have scale with calibrations $\frac{1}{(OY)^2}$ and $\frac{1}{(OX)^2}$, correspondingly, where in the denominators are lengths in the scale of map being used.

Therefore, readings on axes of the plotting board will be

$$Oa_i=\frac{\sin^2\theta_i}{D_i^2\sigma_i^2},\ Ob_i=\frac{\cos^2\theta_i}{D_i^2\sigma_i^2}\ \text{and}\ \sqrt{Oa_iOb_i}=\frac{\sin\theta_i\cos\theta_i}{D_i^2\sigma_i^2}.$$

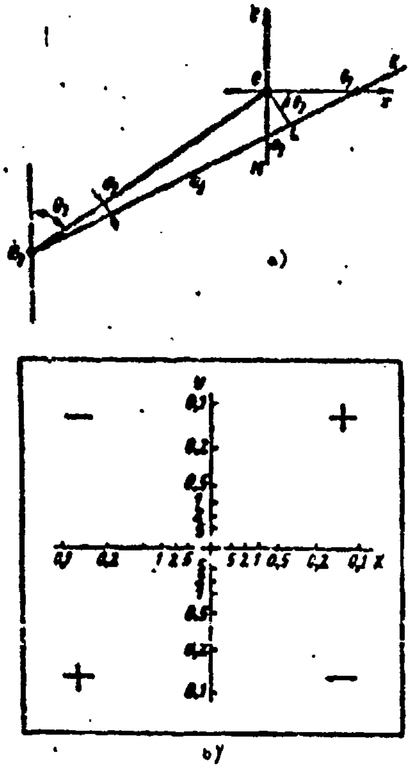


Fig. 11.4. Principle of use of a plotting board for determination parameters of the ellipse of probability: a) diagram of plotting on map; b) form of plotting board.

We add readings on axes of the plotting board for all radio direction finders. Then

$$\sum_{i=1}^{n} Oa_i=C,\ \sum_{i=1}^{n} Ob_i=A\ \text{and}\ \sqrt{\sum_{i=1}^{n} Oa_iOb_i}=B,$$

where A, B, and C correspond to formulas (11.3). Sign for B is shown on the plotting board.

Finding A, B, and C, it is possible to calculate a_0, b_0 and γ from formulas (11.14) and (11.9).

The plotting board is shown in Fig. 11.4b.

It is possible to characterize linear error by the mean quadratic value. In this case it is called by circular error of plotting (R). By definition

$$R=\sqrt{\frac{1}{2\pi}\int_0^{2\pi}\rho^2 d\varphi}, \qquad (11.15)$$

where ρ – radius, from center of ellipse of errors;

φ – angle of radius with initial line of reading, e.g., with the major axis of the ellipse.

Let us rewrite (11.15) in the form

$$R=\sqrt{\frac{1}{2\pi}\left[2\int_0^{\pi/2}(\delta_1^2+\delta_2^2)\,d\varphi\right]}. \qquad (11.15')$$

where δ_1 and δ_2 – conjugate radii of the ellipse of errors.

Let us replace in (11.15') [1.17]

$$\delta_1^2+\delta_2^2=a_0^2+b_0^2$$

and substitute for a_0, b_0 expressions (11.8). Then

$$R=\sqrt{\frac{1}{2}(a_0^2+b_0^2)}=\sqrt{-\ln(1-P_0)}\sqrt{\frac{A+C}{AC-B^2}}. \qquad (11.16)$$

if $-\ln(1 - P_{ell}) = 1$, circular error is

$$R_0 = \sqrt{\frac{A+C}{AC-B^2}}. \quad (11.17)$$

Integral probability of mean quadratic linear error R_0 for $a_0 = b_0$ coincides with integral probability P_{ell} of elliptic error, and according to Fig. 11.3, $P_{ell} = 63.2\%$.

For other ratios $\frac{b_0}{a_0}$ integral probability of circular error R_0 increases somewhat, and as $\frac{b_0}{a_0} \to 0$ it is equal to 68%.

Below there are given values of integral probability of circular error R_{circ} (in %) for various $\frac{R}{R_0}$ for extreme values $\frac{b_0}{a_0} = 1$ and $\frac{b_0}{a_0} = 0$.

Table 11.2. Integral Probability R_{circ} in % Depending upon $\frac{R}{R_0}$ and $\frac{b_0}{a_0}$

$\frac{R}{R_0}$		0,3	0,8	1	1,5	2	2,5	3
P_{cl}	$\frac{b_0}{a_0} = 0$	39	58	68	86	95	99	100
	$\frac{b_0}{a_0} = 1$	20	48	63	90	99	100	

For two radio direction finders from general expression (11.17) for R_0 we have

$$R_0 = 0{,}01745 \frac{\sqrt{\sigma_1^2 D_1^2 + \sigma_2^2 D_2^2}}{\sin \alpha_{12}}, \quad (11.18)$$

where σ_1 and σ_2 are expressed in degrees.

Let us return to expressions (11.2) for coordinates of the center of probability; we shall find a method of calculating them. First we determine the point of intersection of bearings from any two radio direction finders, for instance, $(RDF)_j$ and $(RDF)_k$. Equations of lines of bearings we obtain from Fig. 11.2:

$$\left.\begin{aligned} p_j + x\cos\theta_j - y\sin\theta_j = 0, \\ p_k + x\cos\theta_k - y\sin\theta_k = 0. \end{aligned}\right\} \quad (11.19)$$

Coordinate of points of intersection of bearings from $(RDF)_j$ and $(RDF)_k$ from (11.19) will be:

$$\left.\begin{aligned} x_{jk} &= \frac{p_k \sin\theta_j - p_j \sin\theta_k}{\cos\theta_j \sin\theta_k - \sin\theta_j \cos\theta_k} = \\ &= \frac{p_k \sin\theta_j - p_j \sin\theta_k}{\sin(\theta_k - \theta_j)}, \\ y_{jk} &= \frac{p_k \cos\theta_j - p_j \cos\theta_k}{\sin(\theta_k - \theta_j)}. \end{aligned}\right\} \quad (11.20)$$

Let us introduce for characteristic of the weight of the point of intersection of the j-th and k-th bearings as an estimate of the weight of magnitude m_{jk}, the inverse of the square of the area of the unit ellipse of probability, obtained if the radio station is fixed by two radio direction finders, the j-th and k-th:

$$m_{jk}=\frac{\pi^2}{S^2_{ejk}},$$

where $S_{ell\ jk}$ – area of unit ellipse of probability,

$$S_{ejk}=\pi\frac{E_jE_k}{\sin\alpha_{kj}};$$

α_{kj} – angle of intersection of j-th and k-th bearings, equal to $\theta_k - \theta_j$;

$$m_{jk}=\frac{\sin^2(\theta_k-\theta_j)}{E_j^2E_k^2}. \qquad (11.21)$$

We obtain from formulas (11.20) and (11.21) products $x_{jk}m_{jk}$ and $y_{jk}m_{jk}$:

$$x_{jk}m_{jk}=\frac{1}{E_j^2E_k^2}(p_k\sin\theta_j-p_j\sin\theta_k)\sin(\theta_k-\theta_j),$$
$$y_{jk}m_{jk}=\frac{1}{E_j^2E_k^2}(p_k\cos\theta_j-p_j\cos\theta_k)\sin(\theta_k-\theta_j).$$

It is possible to show that

$$\left.\begin{aligned}\sum_{j=1}^{n}\sum_{k=1}^{n}x_{jk}m_{jk}&=\sum_{j=1}^{n}p_j\frac{B\sin\theta_j-C\cos\theta_j}{E_j^2},\\ \sum_{j=1}^{n}\sum_{k=1}^{n}y_{jk}m_{jk}&=\sum_{j=1}^{n}p_j\frac{A\sin\theta_j-B\cos\theta_j}{E_j^2},\\ \sum_{j=1}^{n}\sum_{k=1}^{n}m_{jk}=\sum_{j=1}^{n}\sum_{k=1}^{n}&\frac{\sin^2(\theta_k-\theta_j)}{E_j^2E_k^2}=AC-B^2,\end{aligned}\right\} \qquad (11.22)$$

where A, B, and C are determined earlier and are expressed by formulas (11.3). Comparing expression (11.2) and (11.22), we arrive at the conclusion that coordinates of the center of probability x_0, y_0 can be calculated by formulas

$$\left.\begin{aligned}x_0&=\frac{\sum_{j=1}^{n}\sum_{k=1}^{n}x_{jk}m_{jk}}{\sum_{j=1}^{n}\sum_{k=1}^{n}m_{jk}},\\ y_0&=\frac{\sum_{j=1}^{n}\sum_{k=1}^{n}y_{jk}m_{jk}}{\sum_{j=1}^{n}\sum_{k=1}^{n}m_{jk}}.\end{aligned}\right\} \qquad (11.23)$$

Formulas (11.23) are analogous to formulas for calculating the center of gravity of masses m_{jk}, located at the intersection points of lines of bearings.

Thus, coordinates of the center of probability x_0, y_0 can be defined as coordinate of the center of gravity of the figure of fixing abcdef ... (Fig. 11.1), at vertices of which are placed masses m_{jk} (11.21), characterizing these vertices.

For two bearings the center of probability coincides with the point of intersection of bearings.

For three radio direction finders masses of vertices of the fixing triangle will be:

$$m_{12}=\frac{\sin^2 \alpha_{12}}{E_1^2E_2^2},\quad m_{13}=\frac{\sin^2 \alpha_{13}}{E_1^2E_3^2},\quad m_{23}=\frac{\sin^2 \alpha_{23}}{E_2^2E_3^2};$$

or, multiplying numerators and denominators by identical factors, it is possible to write them otherwise:

$$m_{12}=\frac{E_3^2 \sin^2 \alpha_{12}}{E_1^2E_2^2E_3^2},\quad m_{13}=\frac{E_2^2 \sin^2 \alpha_{13}}{E_1^2E_2^2E_3^2},\quad m_{23}=\frac{E_1^2 \sin^2 \alpha_{23}}{E_1^2E_2^2E_3^2}. \qquad (11.24)$$

Cancelling identical denominators in expressions (11.24), we obtain for masses

$$m_{12}=D_3^2\sigma_3^2\sin^2\alpha_{12},\quad m_{13}=D_2^2\sigma_2^2\sin^2\alpha_{13},\quad m_{23}=D_1^2\sigma_1^2\sin^2\alpha_{23}. \qquad (11.25)$$

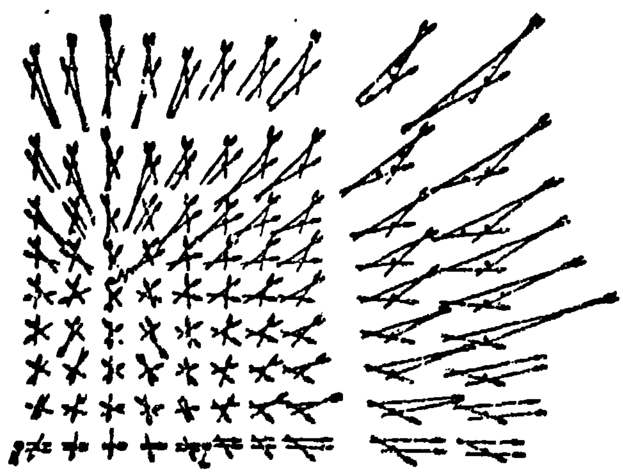


Fig. 11.5. Construction of centers of probability with three radio direction finders.

Graphic determination of the center of probability for three bearings consists in finding the point of intersection of straight lines, drawn from vertices of the fixing triangle so that opposite sides are divided by these lines into segments, inversely proportional to masses of the vertices, adjacent to sides.

In practice we usually mark the center of probability intuitively, proceeding from the position

that the sharper the angle of intersection of bearings at a vertex and the smaller the product $D\sigma$ for the bearing opposite the vertex, the more one should be removed from this vertex.

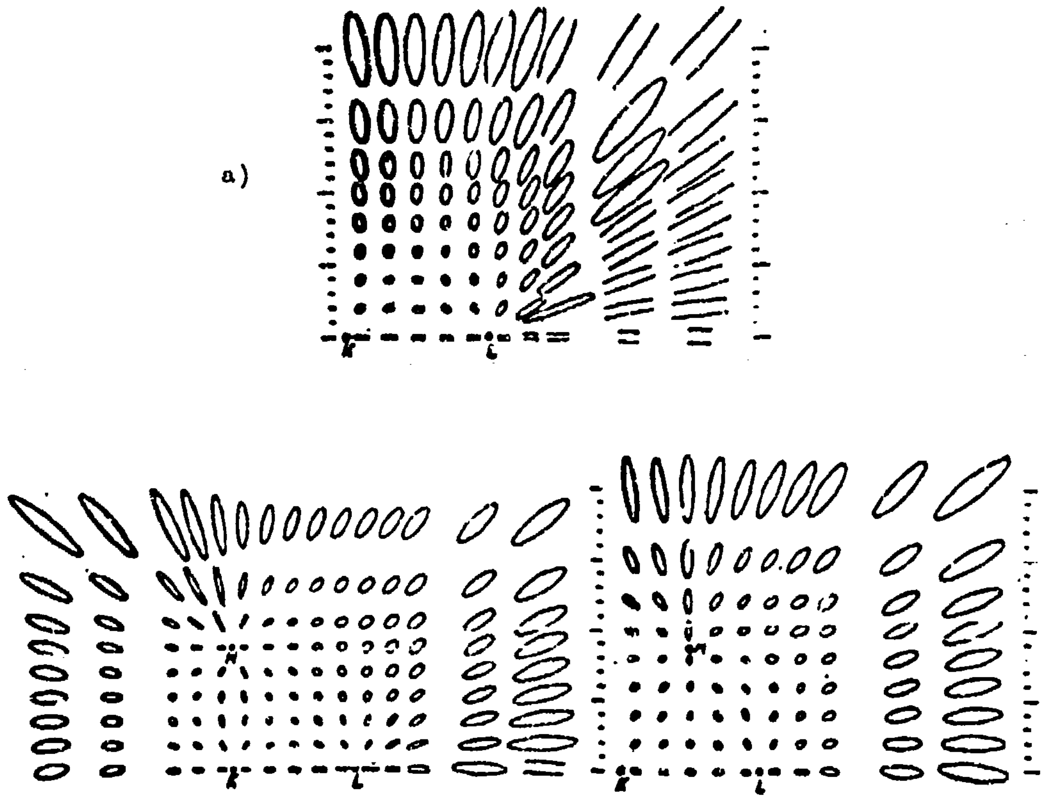


Fig. 11.6. Construction of ellipses of probability for the case of a mean quadratic error of 2^o and $P_{ell} = 50\%$: a) two radio direction finders; b) three radio direction finders at vertices of right-angle triangle; c) three radio direction finder at vertices of an equilateral triangle.

If the center of probability is found and a_0, b_0 and γ, are calculated it is possible to construct the ellipse of probability graphically.

In Fig. 11.5 typical examples of construction of centers of probability are shown for the case of three radio direction finders, located at points K, L, M where $\sigma_1 = \sigma_2 = \sigma_3 = 2^o$.

In Fig. 11.6 are given constructed ellipses of probability for several cases of distribution of radio direction finders [11.6].

Calculations of the center of probability, dimensions of the ellipse of

probability, and orientation of the major axis of the ellipse are given for the case of flat ground, i.e., at distances from the radio direction finders up to 500-800 km. With greater distances it is necessary to allow for curvature of the earth. Equations of lines of bearings must be modified accordingly.

If the lines of bearings are plotted on a map taking into account curvature of the earth and we obtained a fixing figure, then the center of probability and parameters of the ellipse (or circle) of probability can be calculated by the presented theory or by the described graphical method.

11.3. Region, Serviced by Two Radio Direction Finders

Let us determine boundaries of a region, within which linear error of direction finding with required probability will not exceed a given value. The area inside the boundaries of this region is called the working zone of two radio direction finders.

For a given probability the maximum distance at which an object of direction finding can be located from its most probable position, is the major semiaxis of the ellipse of probability.

Thus, determination of the region of direction finding for a given maximum linear error is reduced to finding the region where the major semiaxis of the ellipse does not exceed the given linear error.

For calculation of semiaxes of the ellipse of probability and the angle of orientation of the major axis in the case of two radio direction finders we use formulas (11.10) and (11.11). For simplification of calculations it is conveniently to introduce these parameters (Fig. 11.7): φ — angle between median ON of line $(RDF)_1$, $(RDF)_2$, connecting the direction finders, and the line $(RDF)_1(RDF)_2$; ON — length of the median.

Let us designate $(RDF)_1(RDF)_2 = 2D$, $ON = m$. $(RDF)_1(RDF)_2$ is called the <u>gonimeter base</u> of the two radio direction finders.

Area of triangle $(RDF)_1O(RDF)_2$ is determined by expression

$$\frac{D_1 D_2 \sin\alpha}{2} = \frac{2mD\sin\varphi}{2},$$

whence

$$\sin\alpha = \frac{2mD}{D_1 D_2}\sin\varphi. \qquad (11.26)$$

From triangles $(RDF)_1ON$ and $(RDF)_2ON$ it follows that

$$\left.\begin{aligned} D_1^2 &= m^2 + D^2 + 2mD\cos\varphi, \\ D_2^2 &= m^2 + D^2 - 2mD\cos\varphi. \end{aligned}\right\} \qquad (11.27)$$

Substituting in formulas (11.1) expressions (11.2) and (11.27) and introducing designation $e = \frac{\sigma_2}{\sigma_1}$, we obtain the following equality:

$$a_э = 2\sqrt{-\ln(1-P_э)}\, D e \sigma_1 \times$$

$$\times \sqrt{\frac{\left[\left(\frac{m}{D}\right)^2+1\right]^2 - 4\left(\frac{m}{D}\right)^2 \cos^2\varphi}{T^2 + U - \sqrt{(T^2+U)^2 - 16\left(\frac{m}{D}\right)^2 e^2 \sin^2\varphi}}}, \qquad (11.28)$$

where

$$T^2 = \left[\left(\frac{m}{D}\right)^2 + 1\right](1+e^2),$$

$$U = 2\frac{m}{D}\cos\varphi\,(1-e^2).$$

From expressions (11.28) it follows that when $\varphi = 90°$ the formula for the semimajor axis of the ellipse takes the form

$$a_э = 2\sqrt{-\ln(1-P_э)}\, e\sigma_1 \times$$

$$\times \frac{m^2 + D^2}{\sqrt{(m^2+D^2)(1+e^2) - \sqrt{[(m^2+D^2)(1+e^2)]^2 - 16m^2D^2e^2}}};$$

when $\varphi = 90°$ and $\sigma_1 = \sigma_2 = \sigma$,

$$a_э = \sqrt{-\ln(1-P_э)}\;\sigma\frac{D^2+m^2}{m}, \qquad (11.29)$$

when $D > m$, and

$$a_э = \sqrt{-\ln(1-P_э)}\;\sigma\frac{D^2+m^2}{D}, \qquad (11.30)$$

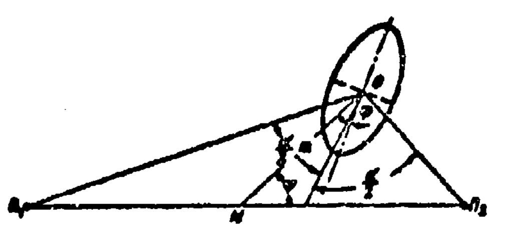


Fig. 11.7. Construction of the ellipse of probability.

when $m > D$. In all these formulas σ, σ_1, and σ_2 are expressed in radians.

When $\varphi = 90$ and $m = D$, i.e., the point of intersection of bearings is at the distance of a semibase (length of the base is equal to 2D) from both radio derection finders, then from (11.29) it follows that

$$a_э = b_э = 2\sqrt{-\ln(1-P_э)}\,\sigma D$$

and the ellipse of probability turns into a circle with radius $2\sqrt{-\ln(1-P_э)}\,\sigma D$. This case corresponds to minimum linear error

$$a_{э\,мин} = 2\sqrt{-\ln(1-P_э)}\,\sigma D = 0{,}035\sqrt{-\ln(1-P_э)}\,\sigma^{\circ} D.$$

When $m = 2D$, i.e., for a point being an identical distance from both radio direction finders, equal to the length of the base,

$$a_0 = 5\sqrt{-\ln(1-P_0)}\,\sigma D = 2.5a_{0\,\text{мин}}.$$

Let us consider methods of constructing the working zone of two radio direction finders.

Let us designate the maximum permissible linear error with probability P_{ell} given by conditions of operation by ΔL.

The condition for finding the working zone of the direction finders will be $a_0 \leq \Delta L$ for the given values of P_{ell}, σ, e, and D.

We introduce in formula (11.28) such a parameter Q that

$$a_0' = 2\sqrt{-\ln(1-P_0)}\,De\sigma Q = \Delta L. \tag{11.31}$$

In this formula

$$Q = \frac{1}{2}\sqrt{\frac{\left[\left(\frac{m}{D}\right)^2+1\right]^2 - 4\left(\frac{m}{D}\right)^2\cos^2\varphi}{\left(\frac{m}{D}\right)^2+1-\sqrt{\left[\left(\frac{m}{D}\right)^2+1\right]^2-4\left(\frac{m}{D}\right)^2\sin^2\varphi}}}, \tag{11.32}$$

when $\sigma_1 = \sigma_2 = \sigma$;

$$Q = \sqrt{\frac{\left[\left(\frac{m}{D}\right)^2+1\right]^2 - 4\left(\frac{m}{D}\right)^2\cos^2\varphi}{T^2+U-\sqrt{(T^2+U)^2-16\left(\frac{m}{D}\right)^2 e^2\sin^2\varphi}}}, \tag{11.33}$$

when $\sigma_1 = \sigma$, $\sigma_2 = e\sigma$.

For given ΔL and P_{ell}, and also σ, e, and D, for the considered radio direction finders it is possible to calculate by (11.31) Q, which depends on φ, e, and $\frac{m}{D}$. Knowing Q and e, it is possible, using (11.32) or (11.33), to calculate the dependence of $\frac{m}{D}$ on φ and to obtain initial data for construction of working zones of the two radio direction finders.

For construction of boundaries of the working zone of direction finders one should:

– to line $(RDF)_1(RDF)_2$, connecting the radio direction finders, pass medians at various angles φ;

– on medians CA plot segments m.

Obtained final points of medians determine boundaries of the working zone of direction finding.

If $e = 1$, i.e., accuracies of both radio direction finders are identical, then boundaries of the zone are symmetric with respect to the perpendicular to center line $(RDF)_1(RDF)_2$; therefore, for angles φ and $180^\circ - \varphi$ we obtain one and the value of $\frac{m}{D}$. If $e \neq 1$, i.e., accuracies of radio direction finders differ, the curve of the working zone is asymmetric. For every value of φ there are two values: $\frac{m}{D}$ and m, bounding on two sides the working zone of direction finding.

To facilitate calculations we give Tables 11.3 and 11.4 of values of Q for various $\frac{m}{2D}$ and φ when $e = 1$ and $e = 2$, calculated by V. V. Shirkov [11.2].

Table 11.3. Value of Parameter Q for Case of Equally-Exact Work of Direction Finders ($e = 1$)

$\frac{m}{2D}$	$\varphi°$					
	15 & 165	30 & 150	45 & 135	60 & 120	75 & 105	90
0,04	13,6	7,08	5,01	4,09	3,66	3,54
0,1	6,69	3,51	2,53	2,10	1,91	1,84
0,2	3,15	1,74	1,32	1,14	1,05	1,03
0,3	1,88	1,17	0,98	0,86	0,82	0,80
0,4	1,19	0,95	0,78	0,82	0,76	0,72
0,5	0,99	0,97	0,92	0,87	0,79	0,71
0,6	1,34	1,14	1,06	0,99	0,01	0,86
0,7	2,00	1,43	1,26	1,16	1,08	1,05
0,8	2,82	1,81	1,52	1,38	1,29	1,26
1,0	4,83	2,79	2,20	1,93	1,81	1,77
1,2	7,22	4,01	3,06	2,64	2,45	2,39
1,4	10,10	5,47	4,09	3,48	3,21	3,13
1,6	13,34	7,15	5,27	4,46	4,09	3,98

During construction of the zone for unequally exact radio direction finders one should consider that $(RDF)_1$ is the more exact radio direction finder and that angles φ should be plotted from $C(RDF)_1$.

With equally exact radio direction finders the maximum range of direction finding, or least error for any distance from the middle of the line of the goniometer base, is obtained along the perpendicular to the middle of the line of the goniometer base.

It is possible to construct boundaries of the region of the working zone, proceeding from obtaining on the boundary of region of the given mean quadratic linear error.

Table 11.4. Value of Parameter Q for Case $e = 2$

$\frac{m}{2D}$	$\varphi°$										
	15	30	45	60	75	90	105	120	135	150	165
0,05	22,80	11,80	8,30	6,70	5,89	5,60	5,70	6,28	7,00	10,6	20,5
0,1	11,76	6,13	4,33	3,52	3,11	2,93	2,91	3,12	3,66	4,98	9,32
0,2	6,18	3,25	2,42	2,03	1,81	1,67	1,59	1,60	1,74	2,18	3,89
0,3	3,73	2,32	1,89	1,67	1,51	1,37	1,24	1,16	1,16	1,32	2,04
0,4	2,33	1,93	1,75	1,62	1,48	1,30	1,17	1,00	0,94	0,99	1,21
0,5	1,95	1,93	1,85	1,73	1,59	1,41	1,22	1,00	0,92	0,97	0,93
0,6	2,67	2,26	2,09	1,95	1,78	1,64	1,39	1,19	1,10	1,16	1,36
0,7	3,97	2,79	2,47	2,26	2,06	1,85	1,64	1,47	1,40	1,52	2,09
0,8	5,63	3,54	2,96	2,64	2,38	2,16	1,95	1,80	1,79	2,00	3,05
1,0	9,15	5,34	4,16	3,58	3,18	2,93	2,73	2,64	2,75	3,29	5,53
1,2	13,45	7,52	5,65	4,75	4,26	3,91	3,72	3,54	3,99	5,00	8,72
1,4	18,73	10,17	7,27	6,07	5,46	5,00	4,80	4,90	5,50	7,06	12,60
1,6	24,33	12,96	9,40	7,76	6,88	6,40	6,23	6,43	7,25	9,49	17,29

For mean quadratic linear error of two radio direction finders we obtained

$$R_0 = \frac{\sqrt{D_1^2\sigma_1^2 + D_2^2\sigma_2^2}}{\sin\alpha_{12}} \ (\sigma_1 \text{ and } \sigma_2 \text{ in radians}) =$$
$$= \frac{0.0175}{\sin\alpha_{12}}\sqrt{D_1^2\sigma_1^2 + D_2^2\sigma_2^2} \ (\sigma_1 \text{ and } \sigma_2 \text{ in degrees}).$$

If we express R_0 depending upon m and φ (see Fig. 11.7), for $\sigma_1 = \sigma_2 = \sigma$ the formula for R_0 will take form [11.3]

$$R_0 = \frac{\sqrt{2}\sigma D}{\sin\varphi}\sqrt{\left[\left(\frac{m}{2D}+\frac{D}{2m}\right)^2 - \cos^2\varphi\right]\left[\left(\frac{m}{D}\right)^2+1\right]}.$$

In order to construct the boundary of the zone of direction finding, for which mean quadratic linear error R_0 will not exceed a given value ΔL, i.e., $R_0 \leq \Delta L$, we use the series of curves of Fig. 11.8, on which there are depicted dependences $\frac{R_0}{2\sigma D} = f(\frac{m}{2D})$ for various values of φ. Minimum error ΔL_{min} is obtained at point $\varphi = 90^\circ$, $m = 0.7D$, where $\alpha_{12} = 109^\circ$, $\Delta L_{min} = 0.032\sigma D$, σ in degrees.

Frequently for the line of the working zone of direction finding we take two circles with radius 2D, passing through direction finders $(RDF)_1$ and $(RDF)_2$. At all points of these circles bearings cross at angles 30° (external circle) and 150° (internal circle).

Calculation shows that linear error for boundaries of the first circumference in its central part is approximately equal to $7a_{0\ min}$; linear error for boundaries of the second circle is equal to $2a_{0\ min}$.

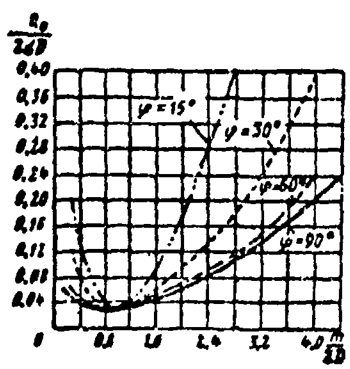


Fig. 11.8. Dependences of $\frac{R_0}{2\sigma D}$ on $\frac{m}{D}$ for various values of φ.

In Table 11.5 there are calculated linear errors for the central part of circles of various angles of intersection of bearings.

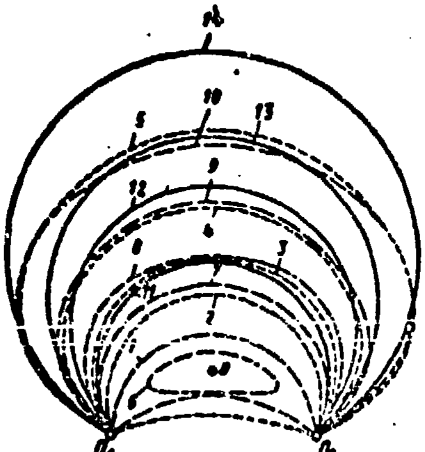
In Fig. 11.9 there are constructed for two radio direction finders contours of zones of direction finding on the boundary of which there is observed constant maximum error (a_0 – semimajor axis of the ellipse of probability) or constant mean quadratic linear error (R_0 – circular error). On the same figure there are drawn neighborhoods of constant angles of intersection of bearings (α_{12} = const).

In the described calculations it was assumed that mean quadratic errors of the radio direction finder kept constant within limits of the whole zone of direction

Table 11.5. Linear Errors $\frac{a_0}{a_{0\,min}}$ for Various Angles of Intersection of Bearings

Angle of intersection of bearings, degrees	90	60 and 150	45 and 160	10 and 170	22.5
$a_0/a_{0\,min}$	1	2	3,4	7,5	13

finding. In fact these errors vary with distance, and when we allow for this construction is considerably complicated.



№	$\frac{R_0}{\sigma D}$	$\frac{a_0}{\sigma D}$	α_{12}°
0	0,052		
1	0,035	0,039	
2		0,052	
3		0,07	
4		0,104	
5		0,176	
6	0,035		
7	0,05		
8	0,064		
9	[illegible]		
10	0,128		
11			60
12			50
13			[illegible]
14	[illegible] / [illegible]		30

Lines of identical maximum errors (a_0)

Lines of identical mean quadratic linear errors (R_0)

Lines of identical angles of intersection of bearings (α_{12})

Fig. 11.9. Zones of radio direction finding for two radio direction finders.

If direction finding is carried out on medium and long waves, the mean quadratic errors are kept approximately constant for the whole working zone. There are limiting distances, beyond which direction finding becomes unreliable. Therefore, after construction of the working zone by mean values of mean quadratic error one should sketch around each radio direction finder maximum regions of direction finding, beyond which direction finding becomes unsatisfactory, and thus cut off areas which are beyond the boundaries of these regions.

With direction finding of short waves is possible to speak of varying angular mean quadratic error at various distances from the radio direction finder. Most simple will be the following characteristic of direction finding:

— at a certain radius r direction finding is impossible (zone of silence);

— at distances from r to R_1 radio direction finders work with reduced accuracy characterized by σ' (zone of steeply incident waves);

— at distances from R_1 to R_2 radio direction finders work with normal accuracy, characterized by σ''.

With such assumptions for determination of the region covered by two radio direction finders it is necessary to construct zones of direction finding for the following conditions:

1) both radio direction finder work with identical accuracy σ';

2) both radio direction finder work with identical accuracy σ'';

3) first direction finder has accuracy, characterized by σ'; and the second, characterized by σ'';

4) first radio direction finder has accuracy σ''; and the second, accuracy σ'.

Furthermore, it is necessary to construct circles with radius r and R_2, separating zones, within which direction finding is impossible.

As a result of such construction we obtain a region of complex outline, sometimes embracing several sections, not interconnected. Individual curves of arcs, bounding the region, intersect at acute or obtuse angles, i.e., there is not a smooth transition of one curve into the other. This is caused by the fact that there is allowed intermittent change of conditions of direction finding. In reality conditions of direction finding vary smoothly; this permits us after construction to somewhat round sharp transitions of one curve, bounding the working zone of direction finding, into another.

When situating a group of radio direction finders for servicing a certain region it is necessary to take into account requirements of obtaining permissible linear error and normal passage of radio waves from the regions to radio direction finders. For more on situating of a group of radio direction finders see [11.1].

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Chapter 12.

LAYING OF RADIO BEARINGS ON MAP/[redacted].

12.1. Orientation of radio direction finder.

Ground radio direction finding stations are establish/installed and orient so that the reading would be obtained equal to zero, when the oriented station is located in the direction of true north. In any other direction the bearing will give the angle of this direction, counting from true direction north - south clockwise, i.e., the true bearing. before the separator of its bearing it is necessary to correct to the value of radio deviation which is taken from tables.

During the installation of stationary radio direction finders, they use precise geodetic instruments. Mobile ground-based radio direction finders it is required to orient after each installation. In appendix IV, are given the easiest methods of determining the direction of true north.

In mobile station (ship, aircraft) the scale of direction finder is oriented so that the null reading is obtained in X direction and readings are conducted clockwise (heading/course angle). For a separator on map/chart, it is necessary bearing to convert for true, i.e., counted off from the direction of the true meridian. As can be seen from Fig. 12.1, true geographical radio station bearings (IP), i.e., counted off from the line of true north, are equal to the sum of the counted off radio bearing (q), radio beam deviations (f), the compass heading (KK), compass error (ΔK) and magnetic declination (ΔM)

$$\text{ИП} = q + f + KK + \Delta M + \Delta K.$$

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The obtained thus true bearing gives direction from ship or aircraft in radio station; with the laying of bearing from radio station, one should take reciprocal bearing.

From the given formula it is evident that the accuracy of

bearing depends on the accuracy of compass and accuracy of the determination of declination and deviation. Furthermore, so with running speed is always possible its divergence from course of 1-2° (but sometimes and is more), it is necessary to simultaneously count off the compass heading and radio bearing.

12.2. Short information about map/charts.

For the image of the Earth in plane, are applied diverse cartographical projections. As a result of the globularity of the Earth, its image on plane is always accompanied by the distortions of one or the other geometric elements. Are utilized at present the diverse forms of cartographical projections. Is given below the short information about the most widely used forms of map/charts.

The projection of Mercator is related to the discharge of cylindrical projections and possesses the following properties: is isogonal (conformal), i.e., transmits without distortion angles and the form of small figures on the earth's surface; orthogonal, i.e., the grid of meridians and parallels mutually perpendicular; loxodromic, i.e., the line, which intersects at sphere meridians at constant angle (loxodromic curve) is depicted straight line.

Last/latter property is especially valuable for navigation, since provides the possibility of the laying of the compasss heading in the form of straight lines. The map/charts of Mercator are fundamental navigation charts.

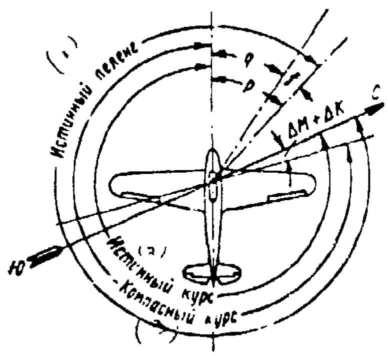

Fig. 12.1. Translation of bearings.

Key: (1). The true bearing. (2). The true course. (3). The compass heading.

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Gnomonic projection - promising, is the projection of sphere from its center on tangential plane. Its most important property, which are inherent only in this form of projection, is orthodromism. All arcs of the great circle on sphere (arc of great circle) are depicted on gnomonic map/chart in the form of straight lines. This property is very important with the separator of radio bearings, since electromagnetic waves are propagated on the shortest distance between two points, i.e., on arc of great circle [1].

FOOTNOTE [1]. Divergences from the shortest path during radiowave propagation are examined in chapter 6 and are related to the errors of direction finding. ENDFOOTNOTE.

However, angles in gnomonic projection are transmitted with distortions.

The value of angular distortions depends on distance of point of contact of tangency (center of the projection of map/chart) and is determined by the formula

$$\operatorname{tg}\alpha' = \cos C \operatorname{tg}\alpha, \qquad (12.1)$$

where α is true azimuth; α' - its image on map/chart; C is the zenith distance of the point in question from point of contact of tangency.

With small zenith distances ($C < 20°$), it is possible to use the approximation formula

$$\Delta = \alpha' - \alpha = -\frac{C^2}{229,2}\sin 2\alpha = -0,353\left(\frac{C_{км}}{1000}\right)^2 \sin 2\alpha. \quad (12.2)$$

At the given point the error depends only on azimuth α. It is possible to construct at the given point azimuth dial/limb taking the correction into account. For zenith distances less than 10° ($C < 1000$ km), error does not exceed 0.5°.

International map/chart of scale 1:1,000,000 is constructed in polyconic projection. On each sheet is reproduced the surface between two meridians, which differ in longitude by 6°, and two parallels, which differ in latitude by 4°. This map/chart within the limits of one sheet can be considered as approximately conformal and great-circle: the distortion of angles does not exceed 7' on middle latitude.

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Due to the properties of polyconic projection during the compound of

four sheets between them are obtained disruptions.

The topograpaic maps of scale from 1:25000 to 1:500,000 are comprised in conformal projection of Gauss. The distortions of angles and divergence of arcs of great circle from straight lines within the limits of map/chart are imperceptible.

Besides the enumerated and entered the universal use map/charts for the laying of radio bearings, can be created the special map/charts, which provide the image of the arcs of great circle, which proceed from the points of the standing of radio direction finder, in the form of straight lines, in this case are retained true angles. This property can be provided not more than at two points of map/chart. A deficiency/lack in such map/charts is the absence of the universality: for each pair of points, is required special map/chart. Frequently is utilized not two, but larger number of direction finders in group.

12.3. Laying of radio bearings on map/chart.

Arc of great circle is depicted on the map/charts of Mercator as curved line. Let in Fig. 12.2 point O designate the position of a

ship on the map/chart of Mercator, point S - the position of the oriented transmitting radio station, line CO - the true meridian, passing through point O, and CS - the true meridian, passing through point S. The straight line OWS at an angle α' is loxodromic curve, arc of great circle (arc of the great circle) will be depicted as certain curve OVS. Radio station is oriented by radio direction finder on vessel O at an angle α. If we lay on map/chart from point O straight line at an angle α to meridian (tangent to arc of great circle OVS), then this straight line passes not through point S, but at certain distance from it.

With the small ranges of laying the curvature of arc of great circle and difference between angles α' and α are so small that it is possible directly on map/chart to run the radio bearing, calculated as is shown above.

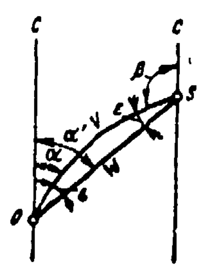

Fig. 12.2. Arc of great circle and loxodromic curve on the map/chart

of Mercator.

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Tables 12.1. Maximum distances for laying without great-circle correction.

(1) Направление на радиостанцию, град				(2) Средняя широта, град													
				70	65	60	55	50	45	40	35	30	25	20	15	10	5
				(3) Предельные расстояния, мили													
0°		180°															
10	170	190	350	56	84	110	140	167	196	224	280	334	420	545	790	1130	2280
20	160	200	340	28	42	56	70	84	98	112	140	168	210	293	370	568	1140
30	150	210	330	20	29	39	49	59	69	79	98	118	147	191	260	395	800
40	140	220	320	15	23	30	38	46	53	61	76	92	114	148	200	307	620
50	130	230	310	13	19	26	32	38	45	51	64	76	96	124	170	258	520
60	120	240	300	12	17	23	29	35	41	47	58	70	87	113	154	235	470
70	110	250	290	11	16	21	26	31	36	42	52	62	78	102	137	210	420
80	100	260	280	10	15	19	24	29	34	38	48	58	72	94	127	195	390
90		270		9	14	18	23	28	32	37	46	56	69	90	122	187	375

Key: (1). Direction on radio station, deg. (2). Middle latitude, deg. (3). Maximum distances, miles.

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Table 12.1 gives the maximum distances at which is possible the laying of radio bearings on the map/chart of Mercator in the form of straight lines without corrections. The difference of angles $\alpha' - \alpha$ at these distances does not exceed 0.3°.

With distances, exceeding indicated by table, it is necessary to consider an angular difference $\varepsilon = \alpha' - \alpha$. This difference, approximately equal to one-half angle of the convergence of meridians δ, he is called great-circle correction. It is designed from the formula

$$\varepsilon = \frac{\delta}{2} = \frac{1}{2} \operatorname{tg} \gamma \sin \varphi_m, \qquad (12.3)$$

where γ is difference in longitude of direction finder and oriented station;

$\varphi_m = \frac{\varphi_0 + \varphi_1}{2}$ - their middle latitude.

Table 12.2 shows the signs of correction for different middle latitudes and the location of transmitting station relative to

direction finder. The sign of correction is given in table such, that correction must be algebraically addded with this sign to the true radio bearing IP in order to obtain loxodromic bearing.

From formula (12.3) it is evident that the great-circle correction the lesser, the lesser difference in longitude and middle latitude. Thus, it is equal to zero, when direction finding is conducted on meridian or when direction finder and the oriented station are equally removed from equator on both sides its.

Tables 12.2. On the calculation of great-circle correction.

(1) Средняя широта	(2) Расположение передающей станции	(3) Знак поправки
(4) Северная	(5) К западу	−
(4) Северная	(6) К востоку	+
(7) Южная	(5) К западу	+
(7) Южная	(6) К востоку	−

Key: (1). Middle latitude. (2). Location of transmitting station. (3). Sign of correction. (4). Northern. (5). To west. (6). To the east. (7). South.

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For determining great-circle correction, it is necessary, at least approximately, to know besides the position of transmitting station also the position of vessel (this approximate position of vessel must be known, also, for determining compass bearing). In the first approximation, it is possible to count that the unknown place lie/rests on the line of loxodromic bearing.

If the obtained location of vessel considerably differs from that which was assuming, it is necessary to convert for a second time bearing, after taking into consideration the new value of the magnetic declination and correction.

During the use of others except mercator map/charts for the separator of radio bearings up to small and average distances, is permissible direct separator without corrections within the limits, indicated in § 12.2. At large distances from point of contact of tangency on gnomonic map/chart, must be taken into account the correction for the distortion of angles (12.1). An international map/chart should be utilized within the limits only of one sheet. At very large distances (on the average more than 1000 nautical miles) the separator on map/charts of any type no longer is not sufficiently precise. The map/chart, which contains both points (direction finder

and the oriented station), extra-fine scale, and determination is obtained imprecise. In these cases they resort on the calculation of, problem of which is to apply to map/chart near the unknown point the small section of the arc of great circle, which corresponds to the obtained bearing. Bearing can be considered as direct/straight on map/chart any type cut. On the basis of the approximate position of the unknown point, we are assigned by its latitude $\varphi_{\kappa 1}$ and we determine the longitude $\gamma_{\kappa 1}$ of the point of intersection of arc of great circle with this latitude by the formula

$$\cos(\gamma_1 - \psi) = \operatorname{tg} h \operatorname{tg} \varphi_{\kappa 1}, \qquad (12.4)$$

where

$$\sin h = \sin \alpha \cos \varphi_{\Gamma};$$

$$\operatorname{tg} \psi = \operatorname{ctg} \alpha \operatorname{cosec} \varphi_{\Gamma};$$

α — истинный пеленг; (1)

φ_{Γ} — широта радиопеленгатора; (2)

γ_{Γ} — долгота радиопеленгатора; (3)

$\gamma_1 = \gamma_{\Gamma} - \gamma_{\kappa 1}$ — разность долгот. (4)

Key: (1). the true bearing. (2). the latitude of radio direction finder. (3). longitude of radio direction finder. (4). difference in longitude.

We further find the slope/inclination of arc of great circle at the particular point:

$$\frac{\Delta\gamma}{\Delta\varphi} = \frac{\operatorname{tg} h}{\cos^2 \varphi_{\kappa 1} \sin(\gamma_1 - \psi)}, \qquad (12.5)$$

On base map, we carry out straight line with the obtained slope/inclination through the point $\varphi_{s1}, \gamma_{s1}$.

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This straight line is the line of position of the oriental object. If there is a second line of position, their intersection determines the position of object and its refined coordinates $\varphi_{s2}, \gamma_{s2}$. The second approach/approximation can be obtained by the repetition of calculation on the basis of the new value of latitude φ_{s2}.

During the use of several bearings, the described construction makes it possible to determine the ellipse of error in accordance with § 11.2.

12.4. Automation of position finding.

The simplest automation of position finding consists in the automatic separator of bearings on map/chart according to the data of radio direction finders. On transparent actual chart area from the

point/items of the arrangement/permutation of radio direction finders, are design/projected the light lines, which correspond to the true radio bearings. In the point of intersection of lines, is located the fixed object. Laying is simplified, if in radio direction finders are utilized the dial instruments these which, transmitted into point/item separators, they serve for control of the design of ray/beams. Is known the use of the equipped thus map/charts with a size/dimension of 100 cm X 100 cm for the maintenance of airfield. The lines of radio bearings can also be design/projected from the screens of the cathode-ray tubes of radio direction finders or their repeaters to large-size map/charts.

By the more advanced method of separator is the automatic separator of bearings electronic method on the screen of the cathode-ray tube on which is plotted/applied the actual chart area. In [12.4], is described the similar tube with a size/dimension of 40 cm X 40 cm with afterglow on which the signals, obtained from six direction-finding point/items, removed from the center of processing up to distances to 60 nautical miles (2 groups on 3 radio direction finders), automatically are processed in line of bearing. Operator can apply the seventh line for determining the course of aircraft for any objective. The images of bearings design/project for the large screen with a size/dimension of 1.5 m X 1.5 m. The design of radio bearings is realized by successive reproduction of lines of bearing

whose directions automatically are established with frequency 400/7 Hz.

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To run bearings in the form of line and to determine the place of object on the intersection of lines is possible only at small distances from radio direction finders, thus far arc of the great circle can be approximated by straight line. For the automatic determination of the coordinates of the fixed object at any distances from direction finders, it is possible to use electronic-computing digital computer. Processing the results of direction finding consists of:

1) coordinate determination of the points of intersection of bearings;

2) the determination of most probable place of object;

3) the calculation of the parameters of the ellipse of probability (size/dimensions of semi-axes and their orientation) for the assigned probability of determination.

As the basis of the solution of the first problem, can be placed

the formula of the calculation of azimuth (12.6). For any two direction-finding point/items are known the azimuths for the objective (bearings) and the coordinates of the points of the standing of radio direction finders. It is required to calculate the coordinates of the point of intersection of the pair of bearings.

Formulas for the calculations of the second and third tasks are given in chapter 11. For their solution it is necessary to feed into the machine estimated average quadratic angular bearing errors.

12.5. Calculation of azimuths.

During the calibration of direction finder, it is required, knowing the coordinates of direction finder (φ_E, γ_E) and transmitter (φ_A, γ_A), to determine the true bearing (azimuth) of the latter for a comparison with radio bearing. This task it is possible to easily solve, utilizing not only map/charts, but also the formulas of spherical trigonometry.

The true bearing α_0 and distance of transmitter C are determined with the aid of the formulas

$$\operatorname{ctg} \alpha_0 = \sin \varphi_E \operatorname{ctg} \gamma - \operatorname{tg} \varphi_A \cos \varphi_E \operatorname{cosec} \gamma, \quad (12.6)$$

$$\sin C = \cos \varphi_A \operatorname{cosec} \alpha_0 \sin \gamma, \quad (12.7)$$

where $\gamma = \gamma_E - \gamma_a$ is difference in longitude.

In the general case of any location of transmitter and direction finder for determining the quadrants of angle α, it is necessary to proceed from the dependences of spherical trigonometry.

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One should add signs to latitudes and longitudes: to north latitude and eastern longitude plus sign; to southern latitude and western longitude minus sign. Then one should determine sign ctg α_0 and use Table 12.3.

During calculation according to formula (12.6) a difference in the terms can pass into sum.

Reciprocal bearing β due to the nonparallelism of meridians is not equal accurately to value α - 180° can be determined by the same formula, but with the corresponding replacement of indices.

It is required to determine simultaneously direct/straight and reciprocal bearings, it is possible to use Napier analogies:

$$\operatorname{tg}\frac{\alpha+\beta}{2}=\frac{\cos\frac{\varphi_1-\varphi_E}{2}}{\sin\frac{\varphi_1+\varphi_E}{2}}\operatorname{ctg}\frac{\gamma}{2}, \tag{12.8}$$

$$\operatorname{tg}\frac{\beta-\alpha}{2}=\frac{\sin\frac{\varphi_1-\varphi_E}{2}}{\cos\frac{\varphi_1+\varphi_E}{2}}\operatorname{ctg}\frac{\gamma}{2},$$

Whence we obtain

$$\alpha=\frac{\alpha+\beta}{2}-\frac{\beta-\alpha}{2}, \tag{12.9}$$

$$\beta=\frac{\alpha+\beta}{2}+\frac{\beta-\alpha}{2}. \tag{12.10}$$

Table 12.3

(4) Положение передатчика относительно пеленгатора	(5) Знак ctg α_0	(6) Квадрант и значение пеленга
(7) К востоку	+	I, α_0
(7) К востоку	—	II, $180^\circ-\alpha_0$
(8) К западу	+	III, $180^\circ+\alpha_0$
(8) К западу	—	IV, $360^\circ-\alpha_0$

Key: (1). Determination. (2). the quadrant of bearing. (3). true. (4). Position of the transmitter of relative direction finder. (5). Sign ctg α_0. (6). Quadrant and the value of bearing. (7). To the east. (8). To west.

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The distance between the direction finder and transmitter D can be determined by the formulas

$$D_{км} = 111C^{\circ},$$

$$\operatorname{tg}\frac{C}{2} = \operatorname{tg}\frac{\varphi_{a} - \varphi_{E}}{2}\,\frac{\sin\frac{\alpha+\beta}{2}}{\sin\frac{\beta-\alpha}{2}} \qquad (12.11)$$

or

$$\operatorname{tg}\frac{C}{2} = \operatorname{ctg}\frac{\varphi_{a} + \varphi_{E}}{2}\,\frac{\cos\frac{\alpha+\beta}{2}}{\cos\frac{\beta-\alpha}{2}}. \qquad (12.12)$$

Formula (12.11) is applied, when $\varphi_{E} + \varphi_{a}$ is small, while formula (12.12) - when $\varphi_{E} - \varphi_{a}$ is small.

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Appendix I.

Calculation of the parameters of the framework.

In the subsequent formulas it is marked:

L - the inductance of framework, μH; C are its capacitance/capacity, pF; S is its area, cm^2; N - turn number; l - the length of wire, cm; g - the space of winding/coil, cm; (d = 2r - diameter of wire, cm;) λ_0 - the natural wavelength of the framework, m; f are frequency, MHz; λ is a wavelength, m; ρ are specific resistor/resistance, Ω/cm; R - a radius of circular framework, cm; a - side of square framework, cm; b is width of framework, cm; μ - magnetic permeability of material (is relative).

1. Inductance of framework. Inductance of the single-turn framework of any form

$$L = 2l\left(\ln\frac{4S}{ld} - 0{,}15\right)10^{-3}.$$

An error in this formula is not above 2-2.5o/o with S/ld > 1000. With other relationship/ratios of size/dimensions, one should use particular formulas for the different figures from which let us give two:

the inductance of the single-turn circular framework

$$L = 4\pi R\left(\ln\frac{8R}{r} - 2\right)10^{-3},$$

the inductance of the single-turn square framework

$$L = 8a\left(\ln\frac{a}{r} + \frac{r}{a} - 0{,}774\right)10^{-3}.$$

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The inductance of the multiturn circular solenoid framework

$$L = 4\pi RN\left\{\left(\ln\frac{R}{r} + 0{,}333\right) + (N-1)\left(\ln\frac{8R}{g} - 2\right) + F\right\}10^{-9} = 2RN(A_1 + B_1)\,10^{-9},$$

where A_1 and B_1 are the values, indicated in Table I.1 and I.2.

Inductance of the multiturn circular framework with the turns, arrange/located (in section) on the apex/vertexes of the regular polygon:

$$L = 4\pi RN\left\{\ln\frac{R}{r} + 0{,}333 + (N-1)\left(\ln\frac{8R}{r_0} - 2\right) - \ln N\right\}10^{-9},$$

where r_0 — a distance of turns from the center of section, ~~see~~ cm

The inductance of the multiturn square framework

$$L = 8aN^2\left\{\ln\frac{a}{(N-1)\,g} + 0{,}2236\,\frac{(N-1)\,g}{a} + 0{,}726 - \left(B + \frac{A}{N}\right)\right\}10^{-9},$$

where A and B are constant, determined from Table I.3 and I.4.

Formula is used both to three-dimensional/space and to the flat/plane framework. In the latter case under a, it is necessary to

understand the average value of the side of the square of turn.

Tables I.1.

$\frac{R}{r}$	400	800	2 000	4 000	10 000	20 000
A_1	39,71	44,04	49,82	54,16	59,94	64,38

Tables I.2.

N	g/R			
	0,002	0,001	0,005	0,02
	Значения B_1 (1)			
2	35,2	30,9	26,5	20,7
6	154,1	132,4	110,4	81,7
10	255,1	216,1	176,6	125,4
14	346,2	289,9	232,5	159,0
18	429,8	355,0	281,5	189,0
22	507,7	415,3	324,2	212,4
26	581,2	417,2	363,2	230,6
30	650,9	523,4	397,7	243,6

Key: (1). Values B_1.

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2. Capacitance/capacity of framework. The capacitance/capacity of the square framework

$$C = C_0 a,$$

where C_0 is the coefficient, determined on Table I.5.

Table I.3.

$\frac{d}{g}$	A	$\frac{a}{g}$	A
1,0	+0,557	0,18	−1,160
0,9	0,452	0,16	1,280
0,8	0,334	0,14	1,410
0,7	0,200	0,12	1,560
0,6	+0,046	0,1	1,750
0,5	−0,136	0,08	1,970
0,4	0,356	0,06	2,26
0,35	0,443	0,04	2,66
0,3	0,647	0,02	−3,36
0,25	0,830		
0,2	1,050		

Tables I.4.

N	B	N	B
1	0	10	0,027
2	0,057	20	0,015
3	0,055	40	0,008
4	0,049	60	0,005
6	0,039	100	0,0033
8	0,032		

Table I.5.

N	1	2	3	4	5	6	Большое
C_0	0,031	0,072	0,102	0,129	0,152	0,167	0,28

Key: (1). Large.

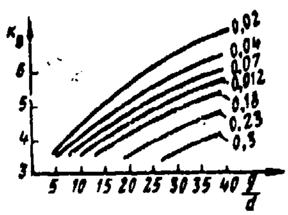

Fig. I.1. Phase constant for calculation λ_0.

3. Natural wavelength of framework: a)

а) $$\lambda_0 = \frac{k_n l}{100},$$

where k_n is coefficient which it is determined for framework from small turn number from Table I.6, and for multiturn square framework from Fig. I.1 with different b/a, designated above curves; where A is coefficient whose values are given in Table I.7.

б) $$\lambda_0 = A\sqrt[3]{\frac{l}{N}}\sqrt{L},$$

4. Resistor/resistance of framework

$$R = R_\Sigma + R_м + R_д + R_з,$$

where R_Σ is radiation resistance, ohm; $R_м$ is resistor/resistance of wires, ohm; $R_д$ is resistor/resistance of dielectric losses, ohm; $R_з$ - resistor/resistance of losses in earth/ground, ohm

$$R_\Sigma = 160\pi^2\frac{h_e^2}{\lambda^2},$$

where h_e - effective height of framework, m;

$$R_м = R_0 k_1 k_2,$$
$$R_0 = \frac{\rho l}{\pi r^2},$$

k_1 - is determined from Table I.8 depending on value $x = 0.14d\sqrt{\frac{\mu f}{\rho}}$. $x = 107d\sqrt{f}$ (for copper).

Tables I.6

Число витков	1	2	4	6
k_n	2,3—2,8	2,75—3	2,7—3,4	3,4—3,6

Tables I.7.

Число витков	4	5	10 и более
A	1,15	1,3	1,5

Key to tables 1.5 (1). Turn number.

Key to tables I.7. (1). Turn number. (2). and more.

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At large values of x k_1 is determined from the formula

$$k_1 = \frac{x\sqrt{2}+1}{4},$$

k_2 - the coefficient, which considers the proximity effect of wires. k_2 it is determined from curve/graph (Fig. I.2) depending on ratio g/d - space of winding/coil toward the diameter of wire. Figure gives two curves on the basis of different sources.

Tables 1.8.

Table 1.8

x	k_1	x	k_1	x	k_1	x	k_1
0	1,000	4,0	1,678	8,8	3,376	19,0	6,974
1,0	1,005	4,4	1,826	9,2	3,517	20,0	7,328
2,0	1,078	4,8	1,971	9,6	3,658	22,0	8,034
2,2	1,111	5,2	2,114	10,0	3,799	24,0	8,741
2,4	1,152	5,6	2,254	11,0	4,151	26,0	9,447
2,6	1,201	6,0	2,394	12,0	4,504	28,0	10,15
2,8	1,256	6,4	2,533	13,0	4,856	30,0	10,88
3,0	1,318	6,8	2,673	14,0	5,209	34,0	12,27
3,2	1,385	7,2	2,813	15,0	5,562	38,0	13,69
3,4	1,456	7,6	2,954	16,0	5,915	42,0	15,10
3,6	1,529	8,0	3,094	17,0	6.268	46,0	16,52
3,8	1,603	8,4	3,235	18,0	6,621	50,0	17,93

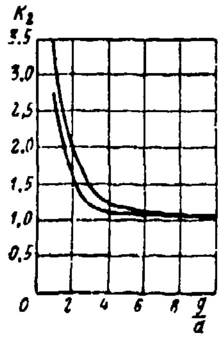

Fig. I.2. On the calculation of the resistor/resistances of the losses of framework.

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For the loop antennas, which consist of very small turn number, the curve/graph of Fig. 1.2 we will not use.

The resistor/resistance of dielectric losses is changed over wide limits depending on the type of construction and cannot be determined by calculation.

The resistor/resistance of the small framework, equivalent to losses in the earth/ground, is usually negligibly small. For the large framework, suspend/hung from masts, this component of resistor/resistance can have prevailing value.

Appendix II.

DERIVATION OF FORMULAS FOR MAGNETIC FIELDS IN MULTIWOUND GONIOMETER.

Equations (3.50) and (3.63) are simplified, since of periodicity

condition of trigonometric functions it follows

$$\sum_{m=1}^{n}\cos\frac{2\pi}{n}pm\cos\frac{2\pi}{n}m\begin{cases}=\frac{n}{2} & \text{при } p=kn+1 \text{ (где } k \text{ может изменяться в пределах } k=0\div\infty) \text{ и при } p=kn-1\ (k=1\div\infty),\\ =0 & \text{при других значениях } p,\end{cases}$$

$$\sum_{m=1}^{n}\cos\frac{2\pi}{n}pm\sin\frac{2\pi}{n}m=0,$$

$$\sum_{m=1}^{n}\sin\frac{2\pi}{n}pm\cos\frac{2\pi}{n}m=0,$$

$$\sum_{m=1}^{n}\sin\frac{2\pi}{n}pm\sin\frac{2\pi}{n}m\begin{cases}=\frac{n}{2} & \text{при } p=kn+1\ (k=0\div\infty),\\ =-\frac{n}{2} & \text{при } p=kn-1\ (k=1\div\infty),\\ =0 & \text{при других } p.\end{cases}$$

Key: (1). with p = kn + 1 (where k can vary within the limits of k = 0 - ∞) and with p = kn - 1 (k = 1 - ∞), at other values of p. (2). with. (3). with other p.

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For a proof one should utilize the transformations:

$$\cos\frac{2\pi}{n}pm\cos\frac{2\pi}{n}m=\frac{1}{2}\left[\cos\frac{2\pi}{n}(p+1)m+\cos\frac{2\pi}{n}(p-1)m\right],$$
$$\sin\frac{2\pi}{n}pm\cos\frac{2\pi}{n}m=\frac{1}{2}\left[\sin\frac{2\pi}{n}(p+1)m+\sin\frac{2\pi}{n}(p-1)m\right],$$
$$\cos\frac{2\pi}{n}pm\sin\frac{2\pi}{n}m=\frac{1}{2}\left[\sin\frac{2\pi}{n}(p+1)m-\sin\frac{2\pi}{n}(p-1)m\right],$$
$$\sin\frac{2\pi}{n}pm\sin\frac{2\pi}{n}m=\frac{1}{2}\left[\cos\frac{2\pi}{n}(p+1)m-\cos\frac{2\pi}{n}(p-1)m\right],$$

and consider that

$$\cos\left[\frac{2\pi}{n}(p+1)m\right]=-\cos\left[\frac{2\pi}{n}(p+1)\left(\frac{n}{2}+m\right)\right]$$

for p ≠ (2n - 1) k, where k can vary from 1 to 1 ∞

$$\cos\left[\frac{2\pi}{n}(p+1)m\right]=\cos\left[\frac{2\pi}{n}(p+1)\left(\frac{n}{2}+m\right)\right]=1$$

for p = (2n - 1) k, where k = 1 — ∞

$$\cos\left[\frac{2\pi}{n}(p-1)m\right]=-\cos\left[\frac{2\pi}{n}(p-1)\left(\frac{n}{2}+m\right)\right]$$

for p ≠ (2n + 1) k, where k = 0 — ∞

$$\cos\left[\frac{2\pi}{n}(p-1)m\right]=\cos\left[\frac{2\pi}{n}(p-1)\left(\frac{n}{2}+m\right)\right]=1$$

for p = (2n - 1) k, where k = 0 — ∞

$$\sin\left[\frac{2\pi}{n}(p\pm 1)m\right] = -\sin\left[\frac{2\pi}{n}(p\pm 1)\left(\frac{n}{2}+m\right)\right] \quad \text{(1) всегда;}$$

$$\sum_{m=1}^{n}\cos\left[\frac{2\pi}{n}(p+1)m\right] = n \quad \text{(2) при } p = kn-1;$$

$$\sum_{m=1}^{n}\cos\left[\frac{2\pi}{n}(p+1)m\right] = 0 \quad \text{(3) в других случаях;}$$

$$\sum_{m=1}^{n}\cos\left[\frac{2\pi}{n}(p-1)m\right] = n \quad \text{при } p = kn+1;$$

$$\sum_{m=1}^{n}\cos\left[\frac{2\pi}{n}(p-1)m\right] = 0 \quad \text{в других случаях;}$$

$$\left.\begin{aligned}\sum_{m=1}^{n}\sin\left[\frac{2\pi}{n}(p+1)m\right] &= 0\\ \sum_{m=1}^{n}\sin\left[\frac{2\pi}{n}(p-1)m\right] &= 0\end{aligned}\right\} \text{(4) во всех случаях.}$$

Key: (1). always. (2). with. (3). in other cases. (4). in all cases.

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Therefore

$$\sum_{p=0}^{\infty} A_p \cos p\theta \sum_{m=1}^{n} \cos \frac{2\pi}{n} pm \cos \frac{2\pi}{n} m + \sum_{p=1}^{\infty} A_p \sin p\theta \times$$

$$\times \sum_{m=1}^{n} \sin \frac{2\pi}{n} pm \cos \frac{2\pi}{n} m =$$

$$= \frac{n}{2} \left[\sum_{k=0}^{\infty} A_{kn+1} \cos (kn+1)\theta + \sum_{k=1}^{\infty} A_{kn-1} \cos (kn-1)\theta \right],$$

$$\sum_{p=0}^{\infty} A_p \cos p\theta \sum_{m=1}^{n} \cos \frac{2\pi}{n} pm \sin \frac{2\pi}{n} m +$$

$$+ \sum_{p=0}^{\infty} A_p \sin p\theta \sum_{m=1}^{n} \sin \frac{2\pi}{n} pm \sin \frac{2\pi}{n} m =$$

$$= \frac{n}{2} \left[\sum_{k=0}^{\infty} A_{kn+1} \sin (kn+1)\theta - \sum_{k=1}^{\infty} A_{kn-1} \sin (kn-1)\theta \right].$$

APPENDIX III.

COMMON/GENERAL/TOTAL EXPRESSIONS FOR THE PARAMETERS OF ELLIPTICAL FIELD.

Let us examine the general case of the equation of ellipse, when in (4.7) when sin α and cos α are the composite coefficients:

$$L=(l+jm)\cos\alpha-(n+jp)\sin\alpha=M_2\cos\alpha-M_1\sin\alpha, \quad \text{(III.1)}$$

$$\text{где } M_2=l+jm;\ M_1=n+jp.$$

We convert III.1:

$$L^2=\cos^2\alpha\,(l^2+m^2)+\sin^2\alpha\,(n^2+p^2)-\sin 2\alpha\,(ln+mp). \quad \text{(III.2)}$$

In order to determine the angle of the orientation of the transverse $\alpha_{\text{ннн}}$, let us equate zero derivative of L^2 in terms of α:

$$\frac{\partial(L^2)}{\partial\alpha}=\sin 2\alpha\,(n^2+p^2-l^2-m^2)-\cos 2\alpha\,[2\,(ln+mp)]=0.$$

Solving, we will obtain expression for $\alpha_{\text{ннн}}$:

$$\operatorname{tg} 2\alpha_{\text{ннн}}=\frac{2\,(ln+mp)}{n^2+p^2-l^2-m^2}. \quad \text{(III.3)}$$

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From formula (III.3), utilizing formulas of the transformation of trigonometric functions, we will obtain

$$\left.\begin{aligned}\cos 2\alpha_{\text{ннн}}&=\frac{n^2+p^2-l^2-m^2}{\sqrt{(n^2+p^2-l^2-m^2)^2+4(ln+mp)^2}},\\ \sin 2\alpha_{\text{ннн}}&=\frac{2\,(ln+mp)}{\sqrt{(n^2+p^2-l^2-m^2)^2+4\,(ln+mp)^2}}.\end{aligned}\right\} \quad \text{(III.4)}$$

Let us find from (III.2) the relation of small and major axes of the magnetic field of the goniometer:

$$\frac{A^2}{B^2}=\frac{L^2\ \text{при}\ \alpha=\alpha_{\text{мин}}+90^\circ}{L^2\ \text{при}\ \alpha=\alpha_{\text{мин}}}=$$
$$=\frac{l^2+m^2+n^2+p^2-\sqrt{(n^2+p^2-l^2-m^2)^2+4(ln+mp)^2}}{l^2+m^2+n^2+p^2+\sqrt{(n^2+p^2-l^2-m^2)^2+4(ln+mp)^2}} \tag{III.5}$$

or

$$\frac{A}{B}=\frac{2(mn-lp)}{l^2+m^2+n^2+p^2+\sqrt{(n^2+p^2-l^2-m^2)^2+4(ln+mp)^2}}. \tag{III.6}$$

A simpler characteristic of the ellipticity of field is obtained from (III.6) in the form:

$$\frac{2\left(\frac{A}{B}\right)}{1-\left(\frac{A}{B}\right)^2}=\frac{2(mn-lp)}{\sqrt{(n^2+p^2-l^2-m^2)^2+4(ln+mp)^2}}=$$
$$=\frac{mn-lp}{ln+mp}\sin 2\alpha_{\text{мин}}. \tag{III.7}$$

Formulas (III.3) and (III.7) are used further for different special cases.

Let us demonstrate that in the small ratios A/B there is the equality:

$$\text{tg}\left(\alpha_{\text{мин}} + j\frac{A}{B}\right) = \frac{M_2}{M_1} = \frac{l + jm}{n + jp}.$$

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It is decomposed the left side of the equality

$$\text{tg}\left(\alpha_{\text{мин}} + j\frac{A}{B}\right) = \frac{\sin 2\alpha_{\text{мин}}}{\cos 2\alpha_{\text{мин}} + \text{ch}\, 2\frac{A}{B}} + $$

$$+ j\frac{\text{sh}\, 2\frac{A}{B}}{\cos 2\alpha_{\text{мин}} + \text{ch}\, 2\frac{A}{B}}, \qquad \text{(III.8)}$$

where at the low values A/B

$$\left.\begin{aligned}
&\operatorname{sh} 2\frac{A}{B} \approx \frac{2\frac{A}{B}}{1-\left(\frac{A}{B}\right)^2} = \\
&= \frac{2(mn-lp)}{\sqrt{(n^2+p^2-l^2-m^2)^2+4(ln+mp)^2}}, \\
&\operatorname{ch} 2\frac{A}{B} \approx 1+\frac{1}{2}\left[\frac{\left(2\frac{A}{B}\right)^2}{1-\left(\frac{A}{B}\right)^2}\right] = \frac{1+\left(\frac{A}{B}\right)^2}{1-\left(\frac{A}{B}\right)^2} = \\
&= \frac{l^2+m^2+n^2+p^2}{\sqrt{(n^2+p^2-l^2-m^2)^2+4(ln+mp)^2}}, \\
&\cos 2\alpha_{\text{мкн}} + \operatorname{ch} 2\frac{A}{B} = \frac{2(n^2+p^2)}{\sqrt{(n^2+p^2-l^2-n^2)^2+4(ln+mp)^2}}.
\end{aligned}\right\} \quad \text{(III.9)}$$

Let us substitute expressions (III.4) and (III.9) into formula (III.8), then we obtain

$$\operatorname{tg}\left(\alpha_{\text{мкн}} + j\frac{A}{B}\right) = \frac{ln+mp}{n^2+p^2} + j\frac{mn-lp}{n^2+p^2}. \quad \text{(III.10)}$$

The right side of equality (III.8) can be presented in this form:

$$\frac{l+jm}{n+jp} = \frac{ln+mp}{n^2+p^2} + j\frac{mn-lp}{n^2+p^2}. \quad \text{(III.11)}$$

From expressions (III.10) and (III.11) follows the validity of the equality

$$\mathrm{tg}\left(\alpha_{мин}+j\frac{A}{B}\right)=\frac{M_2}{M_1}.$$

We convert this equality, after substituting $\alpha_{мин}=\theta+\Delta$, where Δ is a reading error of bearing on the minimum:

$$\mathrm{tg}\left[\theta+\left(\Delta+j\frac{A}{B}\right)\right]=\frac{M_2}{M_1}$$

$$\mathrm{tg}\left(\Delta+j\frac{A}{B}\right)\approx\Delta+j\frac{A}{B}=\frac{M_2\cos\theta-M_1\sin\theta}{M_1\cos\theta+M_2\sin\theta}. \quad \text{(III.12)}$$

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Appendix IV.

Determination of the direction of the true meridian.

The direction of the true meridian can be determined by the different methods, examine/considered in practical astronomy and geodesy. We give the easiest methods, which do not require complex computations.

1. Determination of direction of true north in compass, compass and other instruments with magnetic needle. The advantage of this

method is the possibility of its application/use at any time and independent of the conditions of the weather. For obtaining accurate result, it is necessary to dispose of exact value of the magnetic declination for this met. (of one hundred, which not is always possible. In areas of magnetic anomalies, a precise value of the magnetic declination cannot be shown. The magnetic declination can change in the periods of magnetic storms.

2. Determination of direction of true meridian in map/chart. The map/chart, used for determining the direction of the true meridian, must be sufficient large scale, for example 1:100,000 or 1: 50,000. It is first of all necessary to apply on the map/chart a precise site of installation of radio direction finder. Further with the aid of compass or theodolite, they sight any noticeable object/subject (belfry, factory tube etc.), plotted/applied to map/chart. Distance of object/subject to scale of map/chart must be not less than 100-150 mm.

Let the counted off according to instrument angle be equal to α. Is determined direction in object/subject relative to meridian β in map/chart. They turn alidade to angle $\alpha - \beta$ (clockwise, if $\alpha > \beta$) is establish/installed in this direction the landmark, which determines the direction of the true meridian. For an increase in the accuracy, one should repeat the observations of relatively second

object/subject. If is obtained disagreement, it is necessary to find and to remove its reason (most frequently incorrect mark of the position of radio direction finder on map/chart).

3. Determination of direction of true meridian in Polaris. Polaris, entering the constellation Ursa Minor, is the closest star to celestial to the North Pole. It is removed from pole up to the distance, equal approximately 1°, and in its apparent motion is described around pole small circle. Direction in Polaris coincides with the direction of the true meridian only during its lower and upper culminations. The approximate local time of the passage of Polaris through the meridian is given in Table IV.1. For the calculation of daylight saving time, it is necessary to local time (on table) to add the number of time zone, to take away (for eastern longitude) the longitude of place, expressed in time units, and to add unity.

4. Determination of direction of true meridian by the sun. The direction of the true meridian can be determined by the position of the Sun at true noon.

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For determining daylight saving time, which corresponds to apparent

noon, it is necessary 13 hours to add the number of time zone, to take away the longitude of place, expressed in hours and minutes, and to add the correction whose approximate value is given in Table IV.2 (in min).

Hours, the used determinations apparent noon, must be preliminarily checked, and their error must not exceed 10-15 s. For certain time to the calculated noon they induce the tube of theodolite or another similar instrument on the sun [1]. so that cross hairs in tube would divide the visible image of the Sun in half.

FOOTNOTE [1]. For the observation of the Sun necessary to apply the appropriate filter in ocular (can be used smoked glass). ENDFOOTNOTE.

Further during the motion of the Sun, they revolve smoothly alidade with tube so that the intersection of filaments in tube would coincide with the center of the visible image of the Sun. The second observer at this time follows the hours and with the onset of the torque/moment of apparent noon feeds the signal to the first, which stops the displacement of tube.

Tables IV.2.

(1) Дата	(2) Время прохождения Полярной звезды через меридиан, час
(3) 20 января и 20 июля	(9) 6 и 18
(4) 20 февраля и 20 августа	4
(5) 20 марта и 20 сентября	2
(6) 20 апреля и 20 октября	0
(7) 20 мая и 20 ноября	22
(8) 20 июня и 20 декабря	(9) 8 и 20

Key: (1). Date. (2). Transit time of Polaris through the meridian, the hour. (3). 20 January and 20 July. (4). 20 February and 20 August (5). 20 March and 20 September; (6) 20 April and 20 October; (7) 20 May and 20 November. (8). 20 June and on 20 December. (9). and.

Tables IV.2.

Дата	Поправка	Дата	Поправка	Дата	Поправка
1 января	3,1	1 мая	−2,9	1 сентября	0,1
15 января	9,4	15 мая	−3,8	15 сентября	−4,6
1 февраля	13,7	1 июня	−2,4	1 октября	−10,2
15 февраля	14,3	15 июня	0,2	15 октября	−14,1
1 марта	12,5	1 июля	3,5	1 ноября	−16,3
15 марта	9,6	15 июля	3,8	15 ноября	−15,4
1 апреля	4,1	1 августа	6,3	1 декабря	−11,1
15 апреля	0,2	15 августа	4,5	15 декабря	5,7

Key: (1). Date. (2). Correction. (3). January. (4). May. (5). September. (6). February. (7). June. (8). October. (9). March. (10). July. (11). November. (12). April. (13). August. (14). December.

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In the obtained direction is establish/installed the landmark, which determines the direction of meridian.

For an increase in the accuracy of observation it is possible to repeat several times. For example, the first signal they feed after half-hour to apparent noon. The first observer, after discontinuing the displacement of alidade, is record/written the angle, indicated

by instrument (α_1). Further it renews guidance by the Sun before obtaining of the second signal accurately at noon. At this moment it ceases the displacement of alidade and record/writes new angle (α_2). Similarly he is record/written the third reading of the angle (α_3), which must be produced accurately as later than noon (for example, half-hour), as the first reading was produced earlier than noon. The control of accuracy is the coincidence of values $\alpha_1 + \alpha_3/2$ and α_2. For the true direction of meridian, one should take

$$\alpha = \frac{\alpha_1 + \alpha_3 + 2\alpha_2}{4}.$$

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